National Semiconductor

400015

Data Acquisition Linear Devices

Databook

- *Active Filters*
- *Analog Switches/Multiplexers*
- *Analog-to-Digital Converters*
- *Digital-to-Analog Converters*
- *Sample and Hold*
- *Temperature Sensors*
- *Voltage References*
- *Surface Mount*

For information on additional linear devices, please see the General Purpose Linear Devices and Special Purpose Linear Devices Databooks

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Charles E. Sporck
President, Chief Executive Officer
National Semiconductor Corporation

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Charles E. Sporck
President, Chief Executive Officer
National Semiconductor Corporation

Data Acquisition Linear Devices
Databook

1989 Edition

General Information

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Additional Available Linear Devices
Industry Cross Reference Guide by Part Number
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Additional application information is available as specific application notes or completely compiled in the LINEAR APPLICATIONS HANDBOOK. A product cross reference to the specific application note has been provided. This handbook and the 3-volume set of Linear Data Books represent a complete base of information to the National LINEAR product line.

National Semiconductor

Product Status Definitions

Definition of Terms

Data Sheet Identification	Product Status	Definition
Advance Information	Formative or In Design	This data sheet contains the design specifications for product development. Specifications may change in any manner without notice.
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Available Hybrid Products

Device Number	Databook
ADC1210/ADC1211	Data Acquisition Linear Devices
AF100	Data Acquisition Linear Devices
AF150	Data Acquisition Linear Devices
AF151	Data Acquisition Linear Devices
AH0014/AH0015/AH0019	Data Acquisition Linear Devices
DH0006	Individual Datasheet
DH0008	Individual Datasheet
DH0011	Individual Datasheet
DH0034	Individual Datasheet
DH0035	Individual Datasheet
DH3467	Individual Datasheet
DH3725	Individual Datasheet
LH0002	General Purpose Linear Devices
LH0003	General Purpose Linear Devices
LH0004	General Purpose Linear Devices
LH0020	General Purpose Linear Devices
LH0021/LH0041	General Purpose Linear Devices
LH0022/LH0042/LH0052	General Purpose Linear Devices
LH0023/LH0043	Data Acquisition Linear Devices
LH0024	General Purpose Linear Devices
LH0032	General Purpose Linear Devices
LH0033/LH0063	General Purpose Linear Devices
LH0036	General Purpose Linear Devices
LH0038	General Purpose Linear Devices
LH0044	General Purpose Linear Devices
LH0045	General Purpose Linear Devices
LH0053	Data Acquisition Linear Devices
LH0061	General Purpose Linear Devices
LH0062	General Purpose Linear Devices
LH0070/LH0071	Data Acquisition Linear Devices
LH0075	General Purpose Linear Devices
LH0076	General Purpose Linear Devices
LH0082	General Purpose Linear Devices
LH0084	General Purpose Linear Devices
LH0086	General Purpose Linear Devices
LH0091	Special Purpose Linear Devices
LH0094	Special Purpose Linear Devices
LH0101	General Purpose Linear Devices

Device Number	Databook
LH1605	General Purpose Linear Devices
LH2101	General Purpose Linear Devices
LH2108/LH2308	General Purpose Linear Devices
LH2110/LH2210/LH2310	General Purpose Linear Devices
LH2111/LH2211/LH2311	General Purpose Linear Devices
LH2422	Special Purpose Linear Devices
LH4001	General Purpose Linear Devices
LH4002	General Purpose Linear Devices
LH4003	General Purpose Linear Devices
LH4004	General Purpose Linear Devices
LH4006	General Purpose Linear Devices
LH4008	General Purpose Linear Devices
LH4009	General Purpose Linear Devices
LH4010	General Purpose Linear Devices
LH4011	General Purpose Linear Devices
LH4012	General Purpose Linear Devices
LH4033/LH4063	General Purpose Linear Devices
LH4101	General Purpose Linear Devices
LH4104	General Purpose Linear Devices
LH4105	General Purpose Linear Devices
LH4106	General Purpose Linear Devices
LH4117	General Purpose Linear Devices
LH4118	General Purpose Linear Devices
LH4124	General Purpose Linear Devices
LH4141	General Purpose Linear Devices
LH4161	General Purpose Linear Devices
LH4162	General Purpose Linear Devices
LH4200	General Purpose Linear Devices
LH4266	Special Purpose Linear Devices
LH4860	Data Acquisition Linear Devices
LH7001	General Purpose Linear Devices
LH7070/LH7071	Data Acquisition Linear Devices
HS7067	General Purpose Linear Devices
HS7107	General Purpose Linear Devices
MH0007	Individual Datasheet

National Semiconductor

Additional Available Linear Devices

Device	Databook
HS7067 7 Amp, Multimode, High Efficiency Switching Regulator	General Purpose Linear Devices
HS7107 7 Amp, Multimode, High Efficiency Switching Regulator	General Purpose Linear Devices
LF111 Voltage Comparator	General Purpose Linear Devices
LF147 Wide Bandwidth Quad JFET Input Operational Amplifiers	General Purpose Linear Devices
LF155 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF156 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF157 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF211 Voltage Comparator	General Purpose Linear Devices
LF255 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF256 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF257 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF311 Voltage Comparator	General Purpose Linear Devices
LF347 Wide Bandwidth Quad JFET Input Operational Amplifiers	General Purpose Linear Devices
LF347B Wide Bandwidth Quad JFET Input Operational Amplifiers	General Purpose Linear Devices
LF351 Wide Bandwidth JFET Input Operational Amplifier	General Purpose Linear Devices
LF353 Wide Bandwidth Dual JFET Input Operational Amplifier	General Purpose Linear Devices
LF355 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF356 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF357 Series Monolithic JFET Input Operational Amplifiers	General Purpose Linear Devices
LF400 Fast Settling JFET Input Operational Amplifier	General Purpose Linear Devices
LF401 Precision Fast Settling JFET Input Operational Amplifier	General Purpose Linear Devices
LF411 Low Offset, Low Drift JFET Input Operational Amplifier	General Purpose Linear Devices
LF412 Low Offset, Low Drift Dual JFET Operational Amplifier	General Purpose Linear Devices
LF441 Low Power JFET Input Operational Amplifier	General Purpose Linear Devices
LF442 Dual Low Power JFET Input Operational Amplifier	General Purpose Linear Devices
LF444 Quad Low Power JFET Input Operational Amplifier	General Purpose Linear Devices
LF451 Wide-Bandwidth JFET Input Operational Amplifier	General Purpose Linear Devices
LF453 Wide-Bandwidth Dual JFET Input Operational Amplifier	General Purpose Linear Devices
LF13741 Monolithic JFET Input Operational Amplifier	General Purpose Linear Devices
LH0002 Current Amplifier	General Purpose Linear Devices
LH0003 Wide Bandwidth Operational Amplifier	General Purpose Linear Devices
LH0004 High Voltage Operational Amplifier	General Purpose Linear Devices
LH0020 High Gain Operational Amplifier	General Purpose Linear Devices
LH0021 1.0-Amp Power Operational Amplifier	General Purpose Linear Devices
LH0022 High Performance FET Operational Amplifier	General Purpose Linear Devices
LH0024 High Slew Rate Operational Amplifier	General Purpose Linear Devices
LH0032 Ultra Fast FET-Input Operational Amplifier	General Purpose Linear Devices
LH0033 Fast Buffer Amplifier	General Purpose Linear Devices
LH0036 Instrumentation Amplifier	General Purpose Linear Devices
LH0038 True Instrumentation Amplifier	General Purpose Linear Devices
LH0041 0.2-Amp Power Operational Amplifier	General Purpose Linear Devices
LH0042 Low Cost FET Operational Amplifier	General Purpose Linear Devices
LH0044 Series Precision Low Noise Operational Amplifiers	General Purpose Linear Devices
LH0045 Two Wire Transmitter	General Purpose Linear Devices

Additional Available Linear Devices (Continued)

Device	Databook
LH0052 Precision FET Operational Amplifier	General Purpose Linear Devices
LH0061 0.5 Amp Wide Band Operational Amplifier	General Purpose Linear Devices
LH0062 High Speed FET Operational Amplifier	General Purpose Linear Devices
LH0063 Fast Buffer Amplifier	General Purpose Linear Devices
LH0075 Positive Precision Programmable Regulator	General Purpose Linear Devices
LH0076 Negative Precision Programmable Regulator	General Purpose Linear Devices
LH0082 Optical Communication Receiver/Amplifier	General Purpose Linear Devices
LH0084 Digitally-Programmable-Gain Instrumentation Amplifier	General Purpose Linear Devices
LH0086 Digitally-Programmable-Gain Amplifier	General Purpose Linear Devices
LH0091 True RMS to DC Converter	Special Purpose Linear Devices
LH0094 Multifunction Converter	Special Purpose Linear Devices
LH0101 Power Operational Amplifier	General Purpose Linear Devices
LH1605 5 Amp, High Efficiency Switching Regulator	General Purpose Linear Devices
LH2101A Dual High Performance Operational Amplifier	General Purpose Linear Devices
LH2108 Dual Super Beta Operational Amplifier	General Purpose Linear Devices
LH2110 Dual Voltage Follower	General Purpose Linear Devices
LH2111 Dual Voltage Comparator	General Purpose Linear Devices
LH2201A Dual High Performance Operational Amplifier	General Purpose Linear Devices
LH2210 Dual Voltage Follower	General Purpose Linear Devices
LH2211 Dual Voltage Comparator	General Purpose Linear Devices
LH2301A Dual High Performance Operational Amplifier	General Purpose Linear Devices
LH2308 Dual Super Beta Operational Amplifier	General Purpose Linear Devices
LH2310 Dual Voltage Follower	General Purpose Linear Devices
LH2311 Dual Voltage Comparator	General Purpose Linear Devices
LH2422 CRT Video Driver Amplifier	Special Purpose Linear Devices
LH4001 Wideband Current Buffer	General Purpose Linear Devices
LH4002 Wideband Video Buffer	General Purpose Linear Devices
LH4003 Precision RF Closed Loop Buffer	General Purpose Linear Devices
LH4004 Wideband FET Input Buffer/Amplifier	General Purpose Linear Devices
LH4006 Precision RF Closed Loop Buffer	General Purpose Linear Devices
LH4008 Fast Buffer	General Purpose Linear Devices
LH4009 Fast Buffer	General Purpose Linear Devices
LH4010 Fast FET Buffer	General Purpose Linear Devices
LH4011 Fast Open Loop Buffer	General Purpose Linear Devices
LH4012 Wideband Buffer	General Purpose Linear Devices
LH4033C Fast and Ultra Fast Buffer Amplifiers	General Purpose Linear Devices
LH4063C Fast and Ultra Fast Buffer Amplifiers	General Purpose Linear Devices
LH4101 Wideband High Current Operational Amplifier	General Purpose Linear Devices
LH4104 Fast Settling High Current Operational Amplifier	General Purpose Linear Devices
LH4105 Precision Fast Settling High Current Operational Amplifier	General Purpose Linear Devices
LH4106 ±5V High Speed Operational Amplifier	General Purpose Linear Devices
LH4117 Precision RF Amplifier	General Purpose Linear Devices
LH4118 Low Gain Wide Band RF Amplifier	General Purpose Linear Devices
LH4124C High Slew Rate Operational Amplifier	General Purpose Linear Devices
LH4141C 0.2 Amp Power Operational Amplifier	General Purpose Linear Devices
LH4161 High Speed Operational Amplifier	General Purpose Linear Devices
LH4162 Dual High Speed Operational Amplifier	General Purpose Linear Devices
LH4200 General Purpose GaAs FET Amplifier	General Purpose Linear Devices
LH4266 SPDT RF Switch	Special Purpose Linear Devices
LH7001 Positive/Negative Adjustable Regulator	General Purpose Linear Devices
LM10 Operational Amplifier and Voltage Reference	General Purpose Linear Devices
LM11 Operational Amplifier	General Purpose Linear Devices

Additional Available Linear Devices (Continued)

Device	Databook
LM12(L) 150W Operational Amplifier	General Purpose Linear Devices
LM78G 4-Terminal Adjustable Regulator	General Purpose Linear Devices
LM78L00 Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM78LXX Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM78MG 4-Terminal Positive Regulator	General Purpose Linear Devices
LM78MXX Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM78S40 Universal Switching Regulator Subsystem	General Purpose Linear Devices
LM78XX Series Voltage Regulators	General Purpose Linear Devices
LM79G 4-Terminal Adjustable Regulator	General Purpose Linear Devices
LM79LXXAC Series 3-Terminal Adjustable Negative Regulators	General Purpose Linear Devices
LM79M00 Series 3-Terminal Negative Regulators	General Purpose Linear Devices
LM79MG 4-Terminal Positive Regulator	General Purpose Linear Devices
LM79XX Series 3-Terminal Negative Regulators	General Purpose Linear Devices
LM101A Operational Amplifier	General Purpose Linear Devices
LM102 Voltage Follower	General Purpose Linear Devices
LM104 Negative Regulator	General Purpose Linear Devices
LM105 Voltage Regulator	General Purpose Linear Devices
LM106 Voltage Comparator	General Purpose Linear Devices
LM107 Operational Amplifier	General Purpose Linear Devices
LM108 Operational Amplifier	General Purpose Linear Devices
LM108A Operational Amplifier	General Purpose Linear Devices
LM109 5-Volt Regulator	General Purpose Linear Devices
LM110 Voltage Follower	General Purpose Linear Devices
LM111 Voltage Comparator	General Purpose Linear Devices
LM112 Operational Amplifiers	General Purpose Linear Devices
LM117 3-Terminal Adjustable Regulator	General Purpose Linear Devices
LM117HV 3-Terminal Adjustable Regulator	General Purpose Linear Devices
LM118 Operational Amplifiers	General Purpose Linear Devices
LM119 High Speed Dual Comparator	General Purpose Linear Devices
LM120 Series 3-Terminal Negative Regulator	General Purpose Linear Devices
LM122 Precision Timer	Special Purpose Linear Devices
LM123 3-Amp, 5-Volt Positive Regulator	General Purpose Linear Devices
LM124 Low Power Quad Operational Amplifiers	General Purpose Linear Devices
LM125 Voltage Regulator	General Purpose Linear Devices
LM126 Voltage Regulator	General Purpose Linear Devices
LM133 3-Amp Adjustable Negative Voltage Regulator	General Purpose Linear Devices
LM137 3-Terminal Adjustable Negative Regulator	General Purpose Linear Devices
LM137HV 3-Terminal Adjustable Negative Regulator (High Voltage)	General Purpose Linear Devices
LM138 3-Amp Adjustable Power Regulator	General Purpose Linear Devices
LM139 Low Power Low Offset Voltage Quad Comparator	General Purpose Linear Devices
LM140 Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM140L Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM143 High Voltage Operational Amplifier	General Purpose Linear Devices
LM144 High Voltage, High Slew Rate Operational Amplifier	General Purpose Linear Devices
LM145 Negative 3-Amp Regulator	General Purpose Linear Devices
LM146 Programmable Quad Operational Amplifier	General Purpose Linear Devices
LM148 Quad 741 Operational Amplifiers	General Purpose Linear Devices
LM149 Wide Band Decompensated (A_V(MIN) = 5)	General Purpose Linear Devices
LM150 3-Amp Adjustable Power Regulator	General Purpose Linear Devices
LM158 Low Power Dual Operational Amplifier	General Purpose Linear Devices
LM160 High Speed Differential Comparator	General Purpose Linear Devices
LM161 High Speed Differential Comparator	General Purpose Linear Devices

Additional Available Linear Devices (Continued)

Device	Databook
LM193 Low Power Low Offset Voltage Dual Comparator	General Purpose Linear Devices
LM194 SuperMatch Pair	Special Purpose Linear Devices
LM195 Ultra Reliable Power Transistor	Special Purpose Linear Devices
LM196 10 Amp Adjustable Voltage Regulator	General Purpose Linear Devices
LM201A Operational Amplifier	General Purpose Linear Devices
LM204 Negative Regulator	General Purpose Linear Devices
LM205 Voltage Regulator	General Purpose Linear Devices
LM206 Voltage Comparator	General Purpose Linear Devices
LM207 Operational Amplifier	General Purpose Linear Devices
LM208 Operational Amplifier	General Purpose Linear Devices
LM208A Operational Amplifier	General Purpose Linear Devices
LM210 Voltage Follower	General Purpose Linear Devices
LM211 Voltage Comparator	General Purpose Linear Devices
LM212 Operational Amplifiers	General Purpose Linear Devices
LM218 Operational Amplifiers	General Purpose Linear Devices
LM219 High Speed Dual Comparator	General Purpose Linear Devices
LM221 Precision Preamplifier	General Purpose Linear Devices
LM224 Low Power Quad Operational Amplifiers	General Purpose Linear Devices
LM239 Low Power Low Offset Voltage Quad Comparator	General Purpose Linear Devices
LM246 Programmable Quad Operational Amplifier	General Purpose Linear Devices
LM248 Quad 741 Operational Amplifiers	General Purpose Linear Devices
LM249 Wide Band Decompensated (A_V(MIN) = 5)	General Purpose Linear Devices
LM258 Low Power Dual Operational Amplifier	General Purpose Linear Devices
LM260 High Speed Differential Comparator	General Purpose Linear Devices
LM261 High Speed Differential Comparator	General Purpose Linear Devices
LM293 Low Power Low Offset Voltage Dual Comparator	General Purpose Linear Devices
LM295 Ultra Reliable Power Transistor	Special Purpose Linear Devices
LM301A Operational Amplifier	General Purpose Linear Devices
LM302 Voltage Follower	General Purpose Linear Devices
LM304 Negative Regulator	General Purpose Linear Devices
LM305 Voltage Regulator	General Purpose Linear Devices
LM306 Voltage Comparator	General Purpose Linear Devices
LM307 Operational Amplifier	General Purpose Linear Devices
LM308 Operational Amplifier	General Purpose Linear Devices
LM308A Operational Amplifier	General Purpose Linear Devices
LM309 5-Volt Regulator	General Purpose Linear Devices
LM310 Voltage Follower	General Purpose Linear Devices
LM311 Voltage Comparator	General Purpose Linear Devices
LM312 Operational Amplifiers	General Purpose Linear Devices
LM317 3-Terminal Adjustable Regulator	General Purpose Linear Devices
LM317HV 3-Terminal Adjustable Regulator	General Purpose Linear Devices
LM317L 3-Terminal Adjustable Regulator	General Purpose Linear Devices
LM318 Operational Amplifiers	General Purpose Linear Devices
LM319 High Speed Dual Comparator	General Purpose Linear Devices
LM320 Series 3-Terminal Negative Regulator	General Purpose Linear Devices
LM320L 3-Terminal Negative Regulator	General Purpose Linear Devices
LM321 Precision Preamplifier	General Purpose Linear Devices
LM322 Precision Timer	Special Purpose Linear Devices
LM323 3-Amp, 5-Volt Positive Regulator	General Purpose Linear Devices
LM324 Low Power Quad Operational Amplifiers	General Purpose Linear Devices
LM325 Voltage Regulator	General Purpose Linear Devices
LM326 Voltage Regulator	General Purpose Linear Devices

Additional Available Linear Devices (Continued)

Device	Databook
LM330 3-Terminal Positive Regulator	General Purpose Linear Devices
LM333 3-Amp Adjustable Negative Voltage Regulator	General Purpose Linear Devices
LM337 3-Terminal Adjustable Negative Regulator	General Purpose Linear Devices
LM337HV 3-Terminal Adjustable Negative Regulator (High Voltage)	General Purpose Linear Devices
LM337L 3-Terminal Adjustable Regulator	General Purpose Linear Devices
LM338 3-Amp Adjustable Power Regulator	General Purpose Linear Devices
LM339 Low Power Low Offset Voltage Quad Comparator	General Purpose Linear Devices
LM340 Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM340L Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM341 Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM342 Series 3-Terminal Positive Regulators	General Purpose Linear Devices
LM343 High Voltage Operational Amplifier	General Purpose Linear Devices
LM344 High Voltage, High Slew Rate Operational Amplifier	General Purpose Linear Devices
LM345 Negative 3-Amp Regulator	General Purpose Linear Devices
LM346 Programmable Quad Operational Amplifier	General Purpose Linear Devices
LM348 Quad 741 Operational Amplifiers	General Purpose Linear Devices
LM349 Wide Band Decompensated (A_V(MIN) = 5)	General Purpose Linear Devices
LM350 3-Amp Adjustable Power Regulator	General Purpose Linear Devices
LM358 Low Power Dual Operational Amplifier	General Purpose Linear Devices
LM359 Dual, High Speed, Programmable Current Mode (Norton) Amplifier	General Purpose Linear Devices
LM360 High Speed Differential Comparator	General Purpose Linear Devices
LM361 High Speed Differential Comparator	General Purpose Linear Devices
LM363 Precision Instrumentation Amplifier	General Purpose Linear Devices
LM376 Voltage Regulator	General Purpose Linear Devices
LM380 Audio Power Amplifier	Special Purpose Linear Devices
LM381 Low Noise Dual Preamplifier	Special Purpose Linear Devices
LM382 Low Noise Dual Preamplifier	Special Purpose Linear Devices
LM383 7 Watt Audio Power Amplifier	Special Purpose Linear Devices
LM384 5 Watt Audio Power Amplifier	Special Purpose Linear Devices
LM386 Low Voltage Audio Power Amplifier	Special Purpose Linear Devices
LM387 Low Noise Dual Preamplifier	Special Purpose Linear Devices
LM388 1.5-Watt Audio Power Amplifier	Special Purpose Linear Devices
LM389 Low Voltage Audio Power Amplifier with NPN Transistor Array	Special Purpose Linear Devices
LM390 1 Watt Battery Operated Audio Power Amplifier	Special Purpose Linear Devices
LM391 Audio Power Driver	Special Purpose Linear Devices
LM392 Low Power Operational Amplifier/Voltage Comparator	General Purpose Linear Devices
LM393 Low Power Low Offset Voltage Dual Comparator	General Purpose Linear Devices
LM394 SuperMatch Pair	Special Purpose Linear Devices
LM395 Ultra Reliable Power Transistor	Special Purpose Linear Devices
LM396 10 Amp Adjustable Voltage Regulator	General Purpose Linear Devices
LM431A Adjustable Precision Zener Shunt Regulator	General Purpose Linear Devices
LM494 Pulse Width Modulated Control Circuit	General Purpose Linear Devices
LM555 Timer	Special Purpose Linear Devices
LM555C Timer	Special Purpose Linear Devices
LM556 Dual Timer	Special Purpose Linear Devices
LM556C Dual Timer	Special Purpose Linear Devices
LM565 Phase Locked Loop	Special Purpose Linear Devices
LM565C Phase Locked Loop	Special Purpose Linear Devices
LM566C Voltage Controlled Oscillator	Special Purpose Linear Devices
LM567 Tone Decoder	Special Purpose Linear Devices
LM567C Tone Decoder	Special Purpose Linear Devices

Additional Available Linear Devices (Continued)

Device	Databook
LM592 Differential Video Amplifier	Special Purpose Linear Devices
LM604 4-Channel MUX-Amp	General Purpose Linear Devices
LM607 Precision Operational Amplifier	General Purpose Linear Devices
LM611 Adjustable Micropower Floating Voltage Reference and Single-Supply Operational Amplifier	General Purpose Linear Devices
LM613 Dual Operational Amplifiers, Dual Comparators, and Adjustable Reference	General Purpose Linear Devices
LM614 Quad Operational Amplifier and Adjustable Reference	General Purpose Linear Devices
LM621 Brushless Motor Commutator	Special Purpose Linear Devices
LM627 Precision Operational Amplifiers	General Purpose Linear Devices
LM628 Precision Motion Controller	Special Purpose Linear Devices
LM629 Precision Motion Controller	Special Purpose Linear Devices
LM637 Precision Operational Amplifiers	General Purpose Linear Devices
LM675 Power Operational Amplifier	General Purpose Linear Devices
LM710 Voltage Comparator	General Purpose Linear Devices
LM723 Voltage Regulator	General Purpose Linear Devices
LM725 Operational Amplifier	General Purpose Linear Devices
LM733 Differential Video Amplifier	Special Purpose Linear Devices
LM733C Differential Video Amplifier	Special Purpose Linear Devices
LM741 Operational Amplifier	General Purpose Linear Devices
LM759 Power Operational Amplifier	General Purpose Linear Devices
LM776 Multi-Purpose Programmable Operational Amplifier	General Purpose Linear Devices
LM831 Low Voltage Audio Power Amplifier	Special Purpose Linear Devices
LM832 Dynamic Noise Reduction System DNR	Special Purpose Linear Devices
LM833 Dual Audio Operational Amplifier	General Purpose Linear Devices
LM837 Low Noise Quad Operational Amplifier	General Purpose Linear Devices
LM903 Fluid Level Detector	Special Purpose Linear Devices
LM1035 Dual DC Operated Tone/Volume/Balance Circuit	Special Purpose Linear Devices
LM1036 Dual DC Operated Tone/Volume/Balance Circuit	Special Purpose Linear Devices
LM1037 Dual Four-Channel Analog Switch	Special Purpose Linear Devices
LM1038 Dual Four-Channel Analog Switch	Special Purpose Linear Devices
LM1040 Dual DC Operated Tone/Volume/Balance Circuit with Stereo Enhancement Facility	Special Purpose Linear Devices
LM1042 Fluid Level Detector	Special Purpose Linear Devices
LM1044 Analog Video Switch	Special Purpose Linear Devices
LM1112A Dolby B-Type Noise Reduction Processor	Special Purpose Linear Devices
LM1112B Dolby B-Type Noise Reduction Processor	Special Purpose Linear Devices
LM1112C Dolby B-Type Noise Reduction Processor	Special Purpose Linear Devices
LM1131A Dual Dolby B-Type Noise Reduction Processor	Special Purpose Linear Devices
LM1201 Video Amplifier System	Special Purpose Linear Devices
LM1203 RGB Video Amplifier System	Special Purpose Linear Devices
LM1211 Broadband Demodulator System	Special Purpose Linear Devices
LM1391 Phase-Locked Loop	Special Purpose Linear Devices
LM1414 Dual Differential Voltage Comparator	General Purpose Linear Devices
LM1458 Dual Operational Amplifier	General Purpose Linear Devices
LM1496 Balanced Modulator-Demodulator	Special Purpose Linear Devices
LM1514 Dual Differential Voltage Comparator	General Purpose Linear Devices
LM1524D Regulating Pulse Width Modulator	General Purpose Linear Devices
LM1525A Pulse Width Modulator	General Purpose Linear Devices
LM1527A Pulse Width Modulator	General Purpose Linear Devices
LM1558 Dual Operational Amplifier	General Purpose Linear Devices
LM1575-5.0 Simple Switcher Step-Down Voltage Regulator	General Purpose Linear Devices

Additional Available Linear Devices (Continued)

Device	Databook
LM1578A Switching Regulator	General Purpose Linear Devices
LM1596 Balanced Modulator-Demodulator	Special Purpose Linear Devices
LM1800 Phase-Locked Loop FM Stereo Demodulator	Special Purpose Linear Devices
LM1801 Battery Operated Power Comparator	Special Purpose Linear Devices
LM1812 Ultrasonic Transceiver	Special Purpose Linear Devices
LM1815 Adaptive Sense Amplifier	Special Purpose Linear Devices
LM1818 Electronically Switched Audio Tape System	Special Purpose Linear Devices
LM1819 Air-Core Meter Driver	Special Purpose Linear Devices
LM1823 Video IF Amplifier/PLL Detection System	Special Purpose Linear Devices
LM1830 Fluid Detector	Special Purpose Linear Devices
LM1837 Low Noise Preamplifier for Autoreversing Tape Playback System	Special Purpose Linear Devices
LM1851 Ground Fault Interrupter	Special Purpose Linear Devices
LM1863 AM Radio System for Electronically Tuned Radio	Special Purpose Linear Devices
LM1865 Advanced FM IF System	Special Purpose Linear Devices
LM1866 Low Voltage AM/FM Receiver	Special Purpose Linear Devices
LM1868 AM/FM Radio System	Special Purpose Linear Devices
LM1870 Stereo Demodulator with Blend	Special Purpose Linear Devices
LM1871 RC Encoder/Transmitter	Special Purpose Linear Devices
LM1872 Radio Control Receiver/Decoder	Special Purpose Linear Devices
LM1875 20 Watt Power Audio Amplifier	Special Purpose Linear Devices
LM1877 Dual Power Audio Amplifier	Special Purpose Linear Devices
LM1880 No-Holds Vertical/Horizontal	Special Purpose Linear Devices
LM1881 Video Sync Separator	Special Purpose Linear Devices
LM1884 TV Stereo Decoder	Special Purpose Linear Devices
LM1886 TV Video Matrix D to A	Special Purpose Linear Devices
LM1889 TV Video Modulator	Special Purpose Linear Devices
LM1893 Carrier Current Transceiver	Special Purpose Linear Devices
LM1894 Dynamic Noise Reduction System DNR	Special Purpose Linear Devices
LM1895 Audio Power Amplifier	Special Purpose Linear Devices
LM1896 Dual Power Audio Amplifier	Special Purpose Linear Devices
LM1897 Low Noise Preamplifier for Tape Playback System	Special Purpose Linear Devices
LM1921 1 Amp Industrial Switch	Special Purpose Linear Devices
LM1946 Over/Under Current Limit Diagnostic Circuit	Special Purpose Linear Devices
LM1949 Injector Drive Controller	Special Purpose Linear Devices
LM1951 Solid State 1 Amp Switch	Special Purpose Linear Devices
LM1964 Sensor Interface Amplifier	Special Purpose Linear Devices
LM1965 Advanced FM IF System	Special Purpose Linear Devices
LM2002 8 Watt Audio Power Amplifier	Special Purpose Linear Devices
LM2005 20 Watt Automotive Power Amplifier	Special Purpose Linear Devices
LM2065 Advanced FM IF System	Special Purpose Linear Devices
LM2524D Regulating Pulse Width Modulator	General Purpose Linear Devices
LM2575-5.0 Simple Switcher Step-Down Voltage Regulator	General Purpose Linear Devices
LM2578A Switching Regulator	General Purpose Linear Devices
LM2579 Switching Regulator	General Purpose Linear Devices
LM2877 Dual 4 Watt Power Audio Amplifier	Special Purpose Linear Devices
LM2878 Dual 5 Watt Power Audio Amplifier	Special Purpose Linear Devices
LM2879 Dual 8 Watt Audio Amplifier	Special Purpose Linear Devices
LM2889 TV Video Modulator	Special Purpose Linear Devices
LM2893 Carrier Current Transceiver	Special Purpose Linear Devices
LM2896 Dual Power Audio Amplifier	Special Purpose Linear Devices
LM2900 Quad Amplifier	General Purpose Linear Devices

Additional Available Linear Devices (Continued)

Device	Databook
LM2901 Low Power Low Offset Voltage Quad Comparator	General Purpose Linear Devices
LM2902 Low Power Quad Operational Amplifiers	General Purpose Linear Devices
LM2903 Low Power Low Offset Voltage Dual Comparator	General Purpose Linear Devices
LM2904 Low Power Dual Operational Amplifier	General Purpose Linear Devices
LM2905 Precision Timer	Special Purpose Linear Devices
LM2907 Frequency to Voltage Converter	Special Purpose Linear Devices
LM2917 Frequency to Voltage Converter	Special Purpose Linear Devices
LM2924 Low Power Operational Amplifier/Voltage Comparator	General Purpose Linear Devices
LM2925 Low Dropout Regulator with Delayed Reset	General Purpose Linear Devices
LM2930 3-Terminal Positive Regulator	General Purpose Linear Devices
LM2931 Series Low Drop-Out Regulators	General Purpose Linear Devices
LM2935 Low Dropout Dual Regulator	General Purpose Linear Devices
LM2936 Ultra-Low Quiescent Current 5V Regulator	General Purpose Linear Devices
LM2940 1A Low Dropout Regulator	General Purpose Linear Devices
LM2940C 1A Low Dropout Regulator	General Purpose Linear Devices
LM2941 1A Low Dropout Adjustable Regulator	General Purpose Linear Devices
LM2941C 1A Low Dropout Adjustable Regulator	General Purpose Linear Devices
LM2984C Microprocessor Power Supply System	General Purpose Linear Devices
LM3045 Transistor Array	Special Purpose Linear Devices
LM3046 Transistor Array	Special Purpose Linear Devices
LM3080 Operational Transconductance Amplifier	General Purpose Linear Devices
LM3086 Transistor Array	Special Purpose Linear Devices
LM3089 FM Receiver IF System	Special Purpose Linear Devices
LM3146 High Voltage Transistor Array	Special Purpose Linear Devices
LM3189 FM IF System	Special Purpose Linear Devices
LM3301 Quad Amplifier	General Purpose Linear Devices
LM3302 Low Power Low Offset Voltage Quad Comparator	General Purpose Linear Devices
LM3303 Quad Operational Amplifier	General Purpose Linear Devices
LM3361A Low Voltage/Power Narrow Band FM IF System	Special Purpose Linear Devices
LM3401 Quad Amplifier	General Purpose Linear Devices
LM3403 Quad Operational Amplifier	General Purpose Linear Devices
LM3503 Quad Operational Amplifier	General Purpose Linear Devices
LM3524D Regulating Pulse Width Modulator	General Purpose Linear Devices
LM3525A Pulse Width Modulator	General Purpose Linear Devices
LM3527A Pulse Width Modulator	General Purpose Linear Devices
LM3578A Switching Regulator	General Purpose Linear Devices
LM3820 AM Radio System	Special Purpose Linear Devices
LM3900 Quad Amplifier	General Purpose Linear Devices
LM3905 Precision Timer	Special Purpose Linear Devices
LM3909 LED Flasher/Oscillator	Special Purpose Linear Devices
LM3914 Dot/Bar Display Driver	Special Purpose Linear Devices
LM3915 Dot/Bar Display Driver	Special Purpose Linear Devices
LM3916 Dot/Bar Display Driver	Special Purpose Linear Devices
LM4136 Quad Operational Amplifier	General Purpose Linear Devices
LM4250 Programmable Operational Amplifier	General Purpose Linear Devices
LM4500A High Fidelity FM Stereo Demodulator with Blend	Special Purpose Linear Devices
LM6118 Fast Settling Dual Operational Amplifier	General Purpose Linear Devices
LM6121 High Speed Buffer	General Purpose Linear Devices
LM6125 High Speed Buffer	General Purpose Linear Devices
LM6161 High Speed Operational Amplifier	General Purpose Linear Devices
LM6164 High Speed Operational Amplifier	General Purpose Linear Devices
LM6165 High Speed Operational Amplifier	General Purpose Linear Devices

Additional Available Linear Devices (Continued)

Device	Databook
LM6218 Fast Settling Dual Operational Amplifier	General Purpose Linear Devices
LM6221 High Speed Buffer	General Purpose Linear Devices
LM6225 High Speed Buffer	General Purpose Linear Devices
LM6261 High Speed Operational Amplifier	General Purpose Linear Devices
LM6264 High Speed Operational Amplifier	General Purpose Linear Devices
LM6265 High Speed Operational Amplifier	General Purpose Linear Devices
LM6313 High Speed, High Power Operational Amplifier	General Purpose Linear Devices
LM6321 High Speed Buffer	General Purpose Linear Devices
LM6325 High Speed Buffer	General Purpose Linear Devices
LM6361 High Speed Operational Amplifier	General Purpose Linear Devices
LM6364 High Speed Operational Amplifier	General Purpose Linear Devices
LM6365 High Speed Operational Amplifier	General Purpose Linear Devices
LM7800 Series Voltage Regulators	General Purpose Linear Devices
LM7900 Series 3-Terminal Negative Regulators	General Purpose Linear Devices
LM13080 Programmable Power Operational Amplifier	General Purpose Linear Devices
LM13600 Dual Operational Transconductance Amplifier with Linearizing Diodes and Buffers	General Purpose Linear Devices
LM13700 Dual Operational Transconductance Amplifier with Linearizing Diodes and Buffers	General Purpose Linear Devices
LM18293 Four Channel Push Pull Driver	Special Purpose Linear Devices
LM77000 Power Operational Amplifier	General Purpose Linear Devices
LMC555 CMOS Timer	Special Purpose Linear Devices
LMC567 Low Power Tone Decoder	Special Purpose Linear Devices
LMC568 Low Power Phase-Locked Loop	Special Purpose Linear Devices
LMC660 CMOS Quad Operational Amplifier	General Purpose Linear Devices
LMC662 CMOS Dual Operational Amplifier	General Purpose Linear Devices
LMC669 Auto Zero	General Purpose Linear Devices
LMC835 Digital Controlled Graphic Equalizer	Special Purpose Linear Devices
LMC7660 Switched Capacitor Voltage Converter	General Purpose Linear Devices
LP124 Micropower Quad Operational Amplifier	General Purpose Linear Devices
LP265 Micropower Programmable Quad Comparator	General Purpose Linear Devices
LP311 Voltage Comparator	General Purpose Linear Devices
LP324 Micropower Quad Operational Amplifier	General Purpose Linear Devices
LP339 Ultra-Low Power Quad Comparator	General Purpose Linear Devices
LP365 Micropower Programmable Quad Comparator	General Purpose Linear Devices
LP395 Ultra Reliable Power Transistor	Special Purpose Linear Devices
LP2902 Micropower Quad Operational Amplifier	General Purpose Linear Devices
LP2950 5V Adjustable Micropower Voltage Regulator	General Purpose Linear Devices
LP2951 Adjustable Micropower Voltage Regulator	General Purpose Linear Devices
LPC660 CMOS Quad Operational Amplifier	General Purpose Linear Devices
LPC662 CMOS Dual Operational Amplifier	General Purpose Linear Devices
OP-07 Low Offset, Low Drift Operational Amplifier	General Purpose Linear Devices
TL081CP Wide Bandwidth JFET Input Operational Amplifier	General Purpose Linear Devices
TL082CP Wide Bandwidth Dual JFET Input Operational Amplifier	General Purpose Linear Devices

National Semiconductor

CROSS REFERENCE BY PART NUMBER

A complete interchangeability list of Linear IC's offered by most Integrated Circuit Manufacturers are listed in this section and reference the nearest National Semiconductor Corp. direct replacement or recommended replacement with either an improved or functional replacement. The following notations are appended to assist you in finding the best option.

No reference note ..... "DIRECT REPLACEMENT"

Note (1) ..... "IMPROVED REPLACEMENT" Pin-for-Pin replacement with "SUPERIOR" Electrical Specifications.

Note (2) ..... "FUNCTIONAL REPLACEMENT" Similar device. Consult datasheet to determine the suitability for specific application.

Note (3) ..... "SIMILAR DEVICE" with superior performance. Consult datasheet to determine suitability of the replacement for specific application.

ANALOG DEVICES	NATIONAL	
AD0042	LH0042	(2)
AD101A	LM101A	(1)
AD201A	LM210A	(1)
AD301A	LM301A	(1)
AD3542	LH0042	(2)
AD5035	LH0042	(2)
AD506	LH0022	(2)
AD509	LH0003	(2)
AD521	LH0036	(2)
AD521	LM363	(2)
AD524	LH0038	(2)
AD537	LM331	(2)
AD562	DAC1266	(3)
AD563	DAC1265	(3)
AD565A	DAC1265	
AD566A	DAC1266	
AD567	DAC1230	(2)
AD573	ADC1005	(2)
AD581	LH0070	(1)
AD581	LM581	
AD582	LF398	(2)
AD583	LF198	(3)
AD588	LM369	(2)
AD589M	LM385	(1)
AD589U	LM185	(1)
AD590	LM134	(2)
AD590	LM135	(2)
AD590	LM34	(3)
AD590	LM35	(3)
AD611J	LF411C	(1)
AD611K	LF411AC	(1)
AD614	LH0086	(2)
AD624	LH0038	(2)
AD650	LM331	(2)
AD651	LM331	(2)
AD654	LM331	(2)
AD673	ADC0841	(2)
AD741	LM741	
AD7502	LF13509	(2)
AD7516	CD4066B	(2)
AD7523	DAC0830	(2)
AD7523	DAC0831	(2)
AD7523	DAC0832	(2)
AD7524	DAC0830	(2)
AD7524	DAC0831	(2)
AD7524	DAC0832	(2)
AD7533	DAC1020	
AD7533	DAC1021	
AD7533	DAC1022	
AD7541	DAC1218	(1)
AD7541	DAC1219	(1)
AD7541A	DAC1218	(2)
AD7541A	DAC1219	(2)
AD7542	DAC1208	(2)
AD7542	DAC1209	(2)
AD7542	DAC1210	(2)
AD7545	DAC1208	(2)
AD7545	DAC1209	(2)
AD7545	DAC1210	(2)
AD7548	DAC1230	(2)
AD7548	DAC1231	(2)
AD7548	DAC1232	(2)
AD7552	ADC1220	(2)
AD7552	ADC1225	(2)
AD7571	ADC1005	(2)
AD7571	ADC1025	(2)
AD7575	ADC0820	(2)
AD7576	ADC0820	(2)
AD7578	ADC1205	(2)
AD7578	ADC1225	(2)
AD7820	ADC0820	
ADDAC-08	DAC0800	
ADDAC-08	DAC0801	
ADDAC-08	DAC0802	
ADDAC80	DAC1280+	(1)
ADDAC85	DAC2180+	(1)
ADLH0032	LH0032	(2)
ADLH0033	LH0033	(2)
ADOP07	LM607	(1)

APEX	NATIONAL	
PA01	LH0101	(2)
PA01	LM12	(2)
PA07	LM12	(2)
PA010	LH0101	(2)
PA010	LM12	(2)
PA011	LM12	(2)
PA51	LM12	(2)
PA73	LM12	(2)

BURR-BROWN	NATIONAL	
3507	LM6361	(2)
3533	LH0033	(2)
3542	LH0042	(2)
3550	LM6361	(2)
3551	LM6361	(2)
3553	LH0063	(2)
3554	LH0032	(2)
3571	LM675	(2)
3572	LH0021	(2)
3573	LM675	(2)
3626	LH0036	(2)
3629	LH0038	(2)
3606A6	LH0084	(2)
3606A6	LH0086	(2)
HOS-100	LH0033	(2)
INA102	LH0038	(2)
SHC298A	LF398A	(1)
SHC80	LF398	(2)
SHC85	LF398	(2)

CTS	NATIONAL	
CTS0002	LH002	
CTS0004	LH0004	
CTS0021	LH0021	
CTS0024	LH0024	
CTS0032	LH0032	
CTS0033	LH0033	
CTS0041	LH0041	
CTS0042	LH0042	
CTS2101A	LH2101A	
CTS2111	LH2111	

ELANTEC	NATIONAL	
EHA2500	LM6161	(2)
EHA2502	LM6161	(2)
EHA2505	LM6361	(2)
EHA2510	LM6161	(2)
EHA2512	LM6161	(2)
EHA2515	LM6361	(2)
EHA2520	LM6164	(2)
EHA2522	LM6164	(2)
EHA2525	LM6364	(2)
EHA2600	LM6161	(2)
EHA2602	LM6161	(2)
EHA2605	LM6361	(2)
EHA2620	LM6164	(2)
EHA2622	LM6164	(2)
EHA2625	LM6364	(2)
EL2006	LM6161	(2)
EL2006C	LM6261	(2)
ELH0002	LH0002	(1)
ELH0021	LH0021	(1)
ELH0032	LH0032	(1)
ELH0033	LH0033	(1)
ELH0041	LH0041	(1)
ELH0101	LH0101	(1)

EXAR	NATIONAL	
XR-1001	MF4C-100	(1)
XR-1002	MF4C-50	(1)
XR084	LF347	(1)
XR084M	LF147	(1)
XR1458	LM1458	(1)
XR146	LF146	(1)
XR246	LF246	(1)
XR346	LF346	(1)

HARRIS (Incl. GE/RCA/ INTERSIL)	NATIONAL	
AD7520	DAC1021	
AD7520	DAC1022	
AD7521	DAC1220	
AD7521	DAC1221	
AD7521	DAC1222	
AD7530	DAC1020	(3)
AD7530	DAC1021	(3)
AD7530	DAC1022	(3)
AD7531	DAC1220	
AD7531	DAC1221	
AD7531	DAC1222	
AD7533	DAC1020	
AD7533	DAC1021	
AD7533	DAC1022	
AD7541	DAC1218	
AD7541	DAC1219	
ADC0801	ADC0801	
ADC0802	ADC0802	
ADC0803	ADC0803	
ADC0804	ADC0804	
CA081	LF411M	(2)
CA081A	LF411C	(2)
CA081B	LF411C	(2)
CA081C	TL081C	(2)
CA082	LF412M	(2)
CA082A	LF412C	(2)
CA082B	LF412C	(2)
CA082C	TL082C	(2)
CA084	LF147	(2)
CA084B	LF347B	(2)
CA084C	LF347	(2)
CA124	LM124	(1)
CA139	LM139	(1)
CA139A	LM139A	(1)
CA1458	LM1458	(1)
CA1558	LM1558	(1)
CA158	LM158	(1)
CA158A	LM158A	(1)
CA224	LM224	(1)
CA239	LM239	(1)
CA239A	LM239A	(1)
CA258	LM258	(1)
CA258A	LM258A	(1)
CA301A	LM301A	(1)
CA307	LM307	(1)
CA3105	LM675	(2)
CA311	LM311	(1)
CA324	LM324	(1)
CA3290	LM393	(2)
CA339	LM339	(1)
CA339A	LM339A	(1)
CA3401	LM3401	(1)
CA358	LM358	(1)
CA358A	LM358A	(1)
CA741	LM741	(1)
CA747	LM747	(1)
CA748	LM748	(1)
DG201	LF11201	
DG211	LF13201	
DG212	LF13202	
HA-OP07	LM607	(1)
HA2400	LM604AM	(2)
HA2404	LM604AM	(2)
HA2405	LM604C	(2)
HA2406	LM604C	(2)
HA2500	LM6161	(2)
HA2502	LM6161	(2)
HA2505	LM6361	(2)
HA2510	LM6161	(2)
HA2512	LM6161	(2)
HA2515	LM6361	(2)
HA2520	LH0003	(1)
HA2520	LM6164	(2)
HA2522	LH0003	(1)
HA2522	LM6164	(2)
HA2525	LH0003	(1)
HA2525	LH6364	(2)
HA2530	LH0024	(2)
HA2535	LH0024	(2)
HA2540	LH0032	(2)
HA2541-2	LM6161	(2)
HA2541-5	LM6361	(2)
HA2542	LH0032	(2)
HA2542-2	LM6164	(2)
HA2542-5	LM6164	(2)
HA2600	LM6161	(2)
HA2602	LM6161	(2)
HA2605	LM6361	(2)
HA2620	LM6164	(2)
HA2622	LM6164	(2)
HA2625	LM6364	(2)
HA2640	LH0004	(1)
HA5033	LH0033	(1)
HA5162	LH0062	(2)
HA5180	LH0052	(1)
HF-10	MF10	
HI-201	LF13201	
HI-300	AH5020	(2)
ICH8530	LH0101	(2)
ICL7114	ADC1205	(2)
ICL7114	ADC1225	(2)
ICL7660	LMC7660	(1)
ICL8069	LM313	
ICL8069	LM385-1.2	
IH5009	AH5009	
IH5010	AH5010	
IH5011	AH5011	
IH5012	AH5012	
IH6108	LF13508	
IH6208	LF13509	
LM741	LM741	
μA748	LM748	

HEWLETT-PACKARD	NATIONAL	
HCTL-100	LM628	(3)

HITACHI	NATIONAL	
HA13421A	LM18293	(3)
HA17082	LF353	(1)
HA17082A	LF412	(1)
HA17084	LF347	(1)
HA17084A	LF347B	(1)
HA17094	LM2904	(1)
HA17301	LM3301	(1)
HA17324	LM324	(1)
HA17339	LM339	(1)
HA17358	LM358	(1)
HA17393	LM393	(1)
HA17458	LM1458	(1)
HA17741	LM741	(1)
HA17747	LM747	(1)
HA17901	LM2901	(1)
HA17902	LM2902	(1)
HA17903	LM2903	(1)

LINEAR TECHNOLOGY	NATIONAL	
AD581	LH0070	
AD581	LM581	
LM1009M	LM136-2.5	
LM129	LM129	
LM134	LM134	
LM185	LM185	
LM199	LM199	
LM234	LM234	
LM329	LM329	
LM334	LM334	
LM385	LM385	
LM399	LM399	
LT1001	LM607A	(1)
LT1004C	LM385	
LT1004M	LM185	
LT1009C	LM336-2.5	
LT1019C	LM368	(2)
LT1019M	LM168	(2)
LT1020	LP2951	(3)
LT1021C	LM369	(1)
LT1021M	LM169	(1)
LT1029C	LM336-5.0	
LT1029M	LM136-5.0	
LT1031	LH0070	
LT117A	LM117A	

LT123A	LM123A	
LT138A	LM138A	
LT150A	LM150A	
LT317A	LM317A	
LT323A	LM323A	
LT338A	LM338A	
LT350A	LM350A	
REF-01	LM168	(1)
REF-01	LM368	(1)
SG1524	LM1524D	(1)
SG1525A	LM1525A	(2)
SG1527A	LM1527A	(2)
SG3524	LM3524D	(1)
SG3525A	LM3525A	(2)
SG3527A	LM3527A	(2)

LSI COMPUTER	**NATIONAL**	
LS7261	LM621	(3)
LS7263	LM621	(3)

MICRA	**NATIONAL**
MC0002	LH0002
MC0003	LH0003
MC0004	LH0004
MC0032	LH0032
MC0033	LH0033
MC0041	LH0041
MC0063	LH0063

MICRO POWER	**NATIONAL**
MP108	LM108
MP108A	LM108A
MP155	LF155
MP155A	LF155A
MP156	LF156
MP156A	LF156A
MP157	LF157
MP157A	LF157A
MP208	LM208
MP208A	LM208A
MP2108A	LH2108A
MP308	LM308
MP308A	LM308A
MP355A	LF355A
MP356A	LF356A
MP357A	LF357A
MP5010G	LM185
MP5010G	LM385
MP5010H	LM185
MP5010H	LM385
MP5010L	LM185
MP5010L	LM385
MPOP07	LM607 (1)

MOTOROLA	**NATIONAL**	
AD562A	DAC1266	(2)
AD563A	DAC1265	(2)
DAC-08	DAC0800	
DAC-08	DAC0801	
DAC-08	DAC0802	
LM109H	LM109H	
LM109K	LM109K STEEL	
LM117K	LM117K STEEL	
LM123K	LM123K STEEL	
LM137K	LM137K STEEL	
LM150K	LM150K STEEL	
LM2931	LM2931	
LM309K	LM309K STEEL	
LM317K	LM317K STEEL	
LM323K	LM323K STEEL	
LM337K	LM337K STEEL	
LM350K	LM350K STEEL	
MC1408	DAC0806	
MC1408	DAC0807	
MC1408	DAC0808	
MC1414	LM1414	
LM1436	LM343	(1)
MC14442	ADC0829	(2)
MC14444	ADC0830	(2)
MC145040	ADC0811	(2)
MC145041	ADC0811	
MC1458	LM1458	
MC1496	LM1496	
MC1508	DAC0808	
MC1514	LM1514	
MC1536	LM143	(1)
MC1558	LM1558	
MC1596	LM1596	
MC1709	LM709	
MC1710	LM710	
MC1723	LM723	
MC1723C	LM723C	
MC1741	LM741	
MC1747	LM747	
MC1748	LM748	
MC3301	LM3301	
MC3302	LM3302	
MC3361	LM3361A	(1)
MC34001	LF351	(1)
MC34001A	LF411C	(1)
MC34001B	LF411C	(1)
MC34002	LF353	(1)
MC34002A	LF412A	(1)
MC34002B	LF412C	(1)
MC34004	LF147	(1)
MC34004	LF347	(1)
MC34004B	LF147	(1)
MC34004B	LF347B	(1)
MC3401	LM3401	(1)
MC3410	DAC1020	
MC3412	DAC1265	(2)
MC35001	LF411M	(1)
MC35001A	LF411M	(1)
MC35001B	LF411M	(1)
MC35002	LF412M	(1)
MC35002A	LF412AM	(1)
MC35002B	LF412M	(1)
MC3510	DAC1020	(2)
MC4741	LM348	
MC78LXXACG	LM78LXXACH	
MC78LXXACP	LM78LXXACZ	
MC78LXXCG	LM78LXXCH	
MC78LXXCP	LM78LXXSACZ	
MC78MXXCT	LM341P-XX	
MC78MXXCT	LM342P-XX	
MC78MXXCT	LM78MXXCT	
MC78XXACT	LM340AT-XX	
MC78XXCK	LM78XXCK	
MC78XXCT	LM78XXCT	
MC79LXXACG	LM320H-XX	
MC79LXXACP	LM320LZ-XX	
MC79LXXCP	LM79LXXCZ	
MC79LXXCP	LM79LXXCZ	
MC79MXXAKC	LM320MP-XX	
MC79XXACT	LM320T-XX	
MC79XXAKC	LM320K-XX	
MC79XXCK	LM320K-XX	
MC79XXCK	LM79XXCK	
MC79XXCT	LM79XXCT	

PRECISION MONOLITHIC INC.	**NATIONAL**	
ADC-910	ADC1005	(2)
ADC-910	ADC1025	(2)
AMP-01	LH0038	(2)
BUF-03	LH0033	(1)
DAC-02	DAC1020	(2)
DAC-02	DAC1021	(2)
DAC-02	DAC1022	(2)
DAC-03	DAC1020	(2)
DAC-03	DAC1021	(2)
DAC-03	DAC1022	(2)
DAC-05	DAC1020	(2)
DAC-05	DAC1021	(2)
DAC-05	DAC1022	(2)
DAC-08	DAC0800	
DAC-08	DAC0801	
DAC-08	DAC0802	
DAC-100	DAC1020	(2)
DAC-100	DAC1021	(2)
DAC-100	DAC1022	(2)
DAC-1408	DAC0806	(2)
DAC-1408	DAC0807	(2)
DAC-1408	DAC0808	(2)
DAC-312	DAC1266	
DAC-8012	DAC1208	(2)
DAC-8012	DAC1209	
DAC-8012	DAC1210	
DAC-888	DAC-0830	(2)
DAC-888	DAC0831	(2)
DAC-888	DAC0832	(2)
MUX-08E	LF13508	
MUX-24E	LF13509	
OP-05	LM607	(2)
OP-07	LM607	(1)
OP-15	LF411	(1)
OP-215	LF412	(1)
OP-77	LM607	(1)
PM-108	LM108	
PM-108A	LM108A	
PM-139	LM139	
PM-139A	LM139A	
PM-155	LF155	
PM-155A	LF155A	
PM-156	LF156	
PM-156A	LF156A	
PM-157	LF157	
PM-157A	LF157A	
PM-208	LM208	
PM-208A	LM208A	
PM-2108A	LH2108A	
PM-308	LM308	
PM-308A	LM308A	
PM-339A	LM339A	
PM-355	LF355	
PM-355A	LF355A	
PM-356	LF356	
PM-357	LF357	
PM-357A	LF357A	
PM-725	LM725	
PM-741	LM741	
PM-747	LM747	
PM-7533	DAC1020	
PM-7533	DAC1021	
PM-7533	DAC1022	
PM-7541	DAC1218	
PM-7541	DAC1219	
PM356A	LF356A	
PM420	LM124	(1)

REF-01CJ	LM368-1.0	(1)
REF-02	LM368-5.0	(3)
REF-43	LM368-2.5	(1)
REF-01	LM369	(1)
SW-06B	LF11333	
SW-06F	LF13333	
SW-06G	LF13333	
SW-201B	LF11201	
SW-201F	LF13201	
SW-201G	LF13201	
SW-202B	LF11202	
SW-202F	LF13202	
SW-202G	LF13202	

RAYTHEON	**NATIONAL**	
LP365	LP365	
RC1458	LM1458	
RC1558	LM1558	
RC714	LM607	(1)
RC741	LM741	
RC747	LM747	
REF-01	LM369	(1)
REF-01T	LM368	(1)
REF-02	LM368-5.0	(3)
REF-03	LM368-5.0	(1)

SAMSUNG	**NATIONAL**
KA3524	LM3524D
KA431	LM431
KA78S40	LM78S40
LM741	LM741
MC78LXX	LM78LXX
MC78MXX	LM78MXX
MC78XX	LM78XX
MC79MXX	LM79MXX
MC79XX	LM79XX

SGS-THOMSON	**NATIONAL**	
L123CB	LM723CN	(1)
L293	LM18293	
L4940	LM2940T-5.0	(2)
L4941	LM2940T-5.0	(2)
L4960	LM2579	(2)
L4962	LM2579	(2)
L78MXXCV	LM341P-XX	(2)
L78S05CV	LM323K-5.0	(1)
L78XXACV	LM340AT-XX	
L78XXCT	LM78XXCK	
L78XXCV	LM78XXCT	
L790XXACV	LM320T-XX	
L79XXCT	LM79XXCK	
L79XXCV	LM79XXCT	
LF198	LF198A	(1)
LF298	LF298	
LF398	LF398A	(1)
LM105H	LM105H	(1)
LM109K	LM109K STEEL	(1)
LM117H	LM117H	(1)
LM117K	LF117K	(1)
LM117K	LF117K STEEL	(1)
LM123K	LF123K STEEL	(1)
LM134	LM134	
LM135	LM135	
LM137H	LM137H	(1)
LM137K	LM137K STEEL	(1)
LM138K	LM138K STEEL	(1)
LM234	LM234	
LM235	LM235	
LM2930A	LM2930T-5.0	(1)
LM2931A	LM2931AT-5.0	
LM2935	LM2935	
LM305H	LM305H	(1)
LM309H	LM309H	(1)
LM309K	LM309K STEEL	(1)
LM317H	LM317H	(1)
LM317K	LM317K	(1)
LM317K	LM317K STEEL	(1)
LM317T	LM317T	(1)
LM323K	LM323K STEEL	(1)
LM334	LM334	
LM335	LM335	
LM335A	LM335A	
LM337H	LM337H	(1)
LM337K	LM337K STEEL	(1)
LM338K	LM338K STEEL	(1)
LM748	LM748	
LM7805MK	LM140K-5.0	(1)
SG1524	LM1524D	(1)
SG1525A	LM1525A	(1)
SG1527A	LM1527A	(1)
SG2524	LM2524D	(1)
SG3524	LM3524D	(1)
SG3525A	LM3525A	(1)
SG3527A	LM3527A	(1)
TBC0136	LM336	
TCA3089	LM3089	
TDA2310	LM381	
μA741	LM741	
μA748	LM748	
μA7805CK	LM7805KC	(1)
μA7812CK	LM7812KC	(1)
μA7812MK	LM140K-12	(1)
μA7815CK	LM7815KC	(1)
μA7815MK	LM140K-15	(1)
μA7905CK	LM7905KC	(1)
μA7905MK	LM120K-5.0	(1)
μA7912CK	LM7912KC	(1)
μA7912MK	LM120K-12	(1)
μA7915CK	LM7915KC	(1)
μA7915MK	LM120K-15	(1)

SIEMENS	**NATIONAL**	
TCA365	LH0101	(1)

SIGNETICS	**NATIONAL**	
78LXXACS	LM78LXXACZ	(1)
78LXXADB	LM78XXACH	(1)
78LXXCDB	LM78LXXCH	(1)
78LXXCS	LM78LXXACZ	(1)
78XXCU	LM78XXCT	(1)
78XXDA	LM78XXCK	(1)
79XXCU	LM79XXCT	(1)
79XXDA	LM79XXCK	(1)
ADC0801	ADC0801	
ADC0802	ADC0802	
ADC0803	ADC0803	
ADC0804	ADC0804	
ADC0805	ADC0805	
DAC-08	DAC0800	
DAC-08	DAC0801	
DAC-08	DAC0802	
LF198	LF198	
LF298	LF298	
LF398	LF398	
LM109DB	LM109H	(1)
LM309DA	LM309K	(1)
LM309DB	LM309H	(1)
LM340XXDA	LM340KXX	
LM340XXLL	LM340T-XX	
MC1408	DAC0806	
MC1408	DAC0807	
MC1408	DAC0808	
MC1496N	LM1496N	
MC1508	DAC0808	
MC1596K	LM1596H	
NE4558	LM833	(2)
NE4558D	LM833CM	(2)
NE4558N	LM833CN	(2)
NE5034	ADC0841	(2)
NE5118	DAC0830	(2)
NE529	LM361	(1)
NE532	LM358	(1)
NE5410	DAC1020	(2)
NE5532	LM833	
NE5532N	LM833CN	(2)
NE5532P	LM833CN	(2)
NE555N	LM555CN	
SA532	LM2904	(1)
SA534	LM2902	(1)
SE5118	DAC0830	(2)
SE529	LM161	(1)
SE532	LM158	(1)
SE5410	DAC1020	(2)
SE567	LM567	(2)
μA723CF	LM723CJ	(1)
μA723CL	LM723CH	(1)
μA723CN	LM723CN	(1)
μA723F	LM723J	(1)
μA723L	LM723H	(1)
μA741	LM741	
μA747	LM747	

SILICON GENERAL	**NATIONAL**	
SG101	LM101A	
SG101A	LM101A	
SG104	LM104	
SG105	LM105	
SG107	LM107	
SG109	LM109	
SG117	LM117	
SG1173	LM675	(2)
SG117A	LM117A	
SG117MV	LM117HV	
SG120-XX	LM120-XX	
SG123	LM123	
SG123A	LM123A	
SG124	LM124	
SG137	LM137	
SG138	LM138	
SG138A	LM138A	
SG140-XX	LM140-XX	
SG1436	LM343	(1)
SG150	LM150	
SG150A	LM150A	
SG1524	LM1524D	(1)
SG1524B	LM1524D	
SG1525A	LM1525A	(1)
SG1527A	LM1527A	(1)
SG1536	LM143	(1)
SG201	LM201A	
SG201A	LM201A	
SG204	LM204	
SG205	LM205	
SG207	LM207	
SG224	LM224	
SG2524	LM2524D	(1)
SG2524B	LM2524D	
SG301A	LM301A	
SG304	LM304	
SG305	LM305	
SG307	LM307	
SG309	LM309	
SG317	LM317	

SG3173	LM675	(2)
SG317A	LM317A	
SG317MV	LM317HV	
SG320-XX	LM320-XX	
SG323	LM323	
SG323A	LM323A	
SG324	LM324	
SG337	LM337	
SG338	LM338	
SG338A	LM338A	
SG340-XX	LM340-XX	
SG350	LM350	
SG350A	LM350A	
SG3524	LM3524D	(1)
SG3524B	LM3524D	
SG3525A	LM3525A	(1)
SG3527A	LM3527A	(1)
SG723	LM723	
SG723C	LM723C	
SG741	LM741	
SG78XX	LM140-XX	
SG78XXA	LM140A-XX	(2)
SG78XXAC	LM340A-XX	(2)
SG78XXC	LM78XXC	
SG79XX	LM120-XX	
SG79XXA	LM120-XX	(2)
SG79XXAC	LM320-XX	(2)
SG79XXC	LM79XXC	

SILICONIX	**NATIONAL**	
DG201	LF13201	
DG202	LF13202	
DG211	LF13201	
DG212	LF13202	
DG508	LF13508	
DG509	LF13509	

SPRAGUE	**NATIONAL**	
SG3525A	LM3525A	
SG3527A	LM3527A	
UDN2993B	LM18293	(3)

TELEDYNE	**NATIONAL**
TP0032	LH0032
TP0033	LH0033

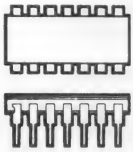
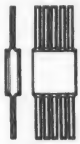
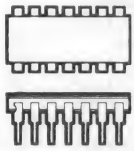
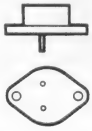
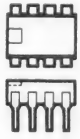
TEXAS INSTRUMENTS	**NATIONAL**
ADC0801	ADC0801
ADC0802	ADC0802
ADC0803	ADC0803
ADC0804	ADC0804
ADC0805	ADC0805
ADC0808	ADC0808
ADC0809	ADC0809
ADC0831	ADC0831
ADC0832	ADC0832

ADC0834	ADC0834	
ADC0838	ADC0838	
LM317KC	LM317T	(1)
RC4558	LM833	
RC4588D	LM833CM	
RV4558D	LM833CM	
TL071	LF351	(1)
TL071A	LF411	(1)
TL071B	LF411	(1)
TL072	LF353	(1)
TL072A	LF412	(1)
TL072B	LF412	(1)
TL074	LF347	(1)
TL074A	LF347B	(1)
TL081	TL081	(1)
TL081A	LF411	(1)
TL081B	LF411	(1)
TL082	TL082	(1)
TL082A	LF412	(1)
TL082B	LF412	(1)
TL084	LF347	(1)
TL084A	LF347B	(1)
TL087	LF411A	(1)
TL088	LF411A	(1)
TL288	LF412A	(1)
TL487N	LM3915N	(2)
TL489N	LM3914N	(2)
TL490N	LM3914N	(2)
TL491N	LM3914N	(2)
TL520	ADC0848	(2)
TL521	ADC0848	(2)
TL522	ADC0848	(2)
TL530	ADC0830B	
TL531	ADC0830C	
TL532	ADC0829B	
TLC532A	ADC0829B	(2)
TL533	ADC0829C	
TLC533A	ADC0829C	(2)
TLC274AC	LMC660AI	(2)
TLC274AI	LMC660AI	(2)
TLC274AM	LMC660AM	(2)
TLC274BC	LMC660AI	(2)
TLC274BI	LMC660AI	(2)
TLC274BM	LMC660AM	(2)
TLC274C	LMC660C	(2)
TLC2741	LMC660AI	(2)
TLC274M	LMC660AM	(2)
TLC540	ADC0811	(2)
TLC541	ADC0811	
TLC549	ADC0831	(2)
TLO61	LF441	(1)
TLO61A	LF441	(1)
TLO61B	LF441A	(1)
TLO62	LF442	(1)
TLO62A	LF442A	(1)
TLO62B	LF442	(1)
TLO64	LF444	(1)
TLO64A	LF444A	(1)
µA709	LM709	
µA723CJ	LM723CJ	
µA723CN	LM723CN	
µA723MJ	LM723J	
µA733CN	LM733CN	
µA741	LM741	
µA747	LM747	
µA78LXXACL	LM78LXXACZ	
µA78MXXCKD	LM78MXXCP	
µA78XXCKC	LM78XXCT	
µA79MXXCKD	LM79MXXCP	
µA79XXCKC	LM79XXCT	

TOSHIBA	**NATIONAL**	
TA7504	LM741	
TA75339	LM2901	(1)
TA75358	LM2904	(1)
TA75393	LM2903	(1)
TA75902	LM2902	(1)

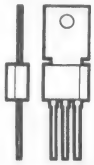
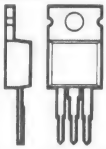
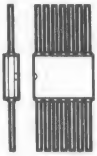
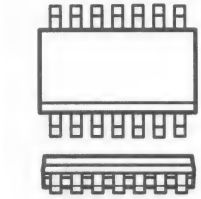
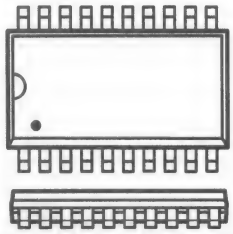
UNITRODE	**NATIONAL**	
L293	LM18293	
UC117	LM117	
UC137	LM137	
UC150	LM150	
UC1524	LM1524D	(1)
UC1524A	LM1524D	(2)
UC1525A	LM1525A	
UC1527A	LM1527A	
UC2524	LM2524D	(1)
UC2524A	LM2524D	(2)
UC317	LM317	
UC337	LM337	
UC350	LM350	
UC3524	LM3524D	(1)
UC3524A	LM3524D	(2)
UC3525A	LM3525A	
UC3527A	LM3527A	
UC494	LM494	(2)
UC78XXACK	LM340AK-XX	(2)
UC78XXAK	LM140AK-XX	(2)
UC78XXCK	LM340K-XX	
UC78XXCK	LM78XXCK	
UC78XXK	LM140K-XX	
UC79XXACK	LM320K-XX	(2)
UC79XXAK	LM120K-XX	(2)
UC79XXCK	LM79XXCK	
UC79XXK	LM120K-XX	(1)

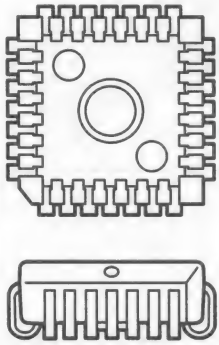
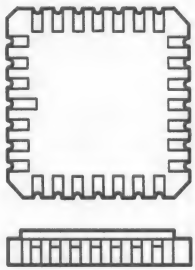
Industry Package Cross-Reference Guide

	NSC	NSC μA	Signetics	Motorola	TI	RCA	Hitachi	NEC	LTC
 4/16 Lead Glass/Metal DIP	D	D	I	L		D	C	D	D
 Glass/Metal Flat Pack	F	F	Q	F	F, S	K	F		Q
TO-99, TO-100, TO-5	H	H	T, K, L, DB	G	L	S*, V1**		A	H
 8-, 14- and 16-Lead Low Temperature Ceramic DIP	J	R, D	F	U	J		G	D	J, J8
 TO-3 (Steel)	K			KS					K
TO-3 (Aluminum)	KC	K	DA	K	K				
 8-, 14- and 16-Lead Plastic DIP	N	T, P	V, A, B	P	P, N	E	P	C	N, N8

*With dual-in-line formed leads

**With radically formed leads

	NSC	NSC μA	Signetics	Motorola	TI	RCA	Hitachi	NEC	LTC
 TO-202 (D-40, Durawatt)	P					KD			
 TO-220 3- & 5-Lead	T	U	U		KC		T	H	T
TO-220 11-, 15- & 23-Lead	T								
 Low Temperature Glass Hermetic Flat Pack	W	F		F	W				
TO-92 (Plastic)	Z	W	S	P	LP			H	Z
 SO (Narrow Body)	M	S	D	D	D	M	MP	G	S
 SO (Wide Body)	WM				DW				

	NSC	NSC μA	Signetics	Motorola	TI	RCA	Hitachi	NEC	LTC
 PCC	V	Q	A	FN	FN	Q	CP	L	
 LCC Leadless Ceramic Chip Carrier	E	L1	G	U	FK/ FG/FH	BJ	CG	K	

Section 1

Active Filters

Section 1 Contents

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Active Filters
Definition of Terms

f_{CLK}: the switched capacitor filter external clock frequency.

f_o: center of frequency of the second order function complex pole pair. f_o is measured at the bandpass output of each ½ MF10, and it is the frequency of the bandpass peak occurrence.

Q: quality factor of the 2nd order function complex pole pair. Q is also measured at the bandpass output of each ½ MF10 and it is the ratio of f_o over the −3 dB bandwidth of the 2nd order bandpass filter. The value of Q is not measured at the lowpass or highpass outputs of the filter, but its value relates to the possible amplitude peaking at the above outputs.

H_{OBP}: the gain in (V/V) of the bandpass output at $f = f_o$.

H_{OLP}: the gain in (V/V) of the lowpass output of each ½ MF10 at $f \rightarrow 0$ Hz.

H_{OHP}: the gain in (V/V) of the highpass output of each ½ MF10 as $f \rightarrow f_{CLK}/2$.

Q_Z: the quality factor of the 2nd order function complex zero pair, if any. (Q_Z is a parameter used when an allpass output is sought and unlike Q it cannot be directly measured).

f_Z: the center frequency of the 2nd order function complex zero pair, if any. If f_Z is different from f_o, and if the Q_Z is quite high it can be observed as a notch frequency at the allpass output.

f_{notch}: the notch frequency observed at the notch output(s) of the MF10.

H_{ON_1}: the notch output gain as $f \rightarrow 0$ Hz.

H_{ON_2}: the notch output gain as $f \rightarrow f_{CLK}/2$.

National Semiconductor

Active Filter Selection Guide

Device #	Type	Function	Max Order	Max Freq Accuracy	Freq Range	Typ. Q Accuracy	Max F x Q
AF100	Universal	Universal	2nd	±1.0%	0.1–10 kHz	±7.5%	50 kHz
AF150	Wideband Universal	Universal	2nd	±1.0%	0.1–100 kHz	±7.5%	200 kHz
AF151	Dual Universal	Universal	4th	±1.0%	0.1–10 kHz	±7.5%	50 kHz
MF10 (S, T)	Universal	Universal	4th	±0.6%	0.1–30 kHz	±2%	200 kHz
MF8 (T)	Bandpass	Chebyshev Butterworth	4th	±1.0%	0.1–20 kHz	±2%	5 MHz
MF6 (S, T)	Lowpass	Butterworth	6th	±1.0%	0.1–20 kHz	N/A	N/A
MF5 (S)	Universal	Universal	2nd	±1.0%	0.1–30 kHz	±6%	200 kHz
MF4 (S)	Lowpass	Butterworth	4th	±0.6%	0.1–20 kHz	N/A	N/A
LMF100	Universal	Universal	4th	±0.6%	0.1–40 kHz	±2%	1.8 MHz
LMF90	Notch	Elliptic	4th	±1%	0.1–30 kHz	N/A	N/A
LMF120	Universal Mask-Programmable	Universal	12th	±1.5%	0.1–100 kHz	±2%	1 MHz

S Surface Mount Available

T Extended Temperature Available

National Semiconductor

AF100 Universal Active Filter

General Description

The AF100 state variable active filter is a general second order lumped RC network. Only four external resistors program the AF100 for specific second order functions. Lowpass, highpass, and bandpass functions are available simultaneously at separate outputs. Notch and allpass functions are available by summing the outputs in the uncommitted output summing amplifier. Higher order systems are realized by cascading AF100 active filters with appropriate programming resistors.

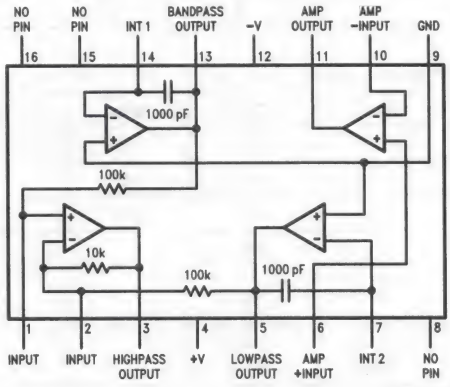
Any of the classical filter configurations, such as Butterworth, Bessel, Cauer, and Chebyshev can be formed.

Features

- Military or commercial specifications
- Independent Q, frequency, gain adjustments
- Low sensitivity to external component variation
- Separate lowpass, highpass, bandpass outputs
- Inputs may be differential, inverting, or non-inverting
- Allpass and notch outputs may be formed using uncommitted amplifier
- Operates to 10 kHz
- Q range to 500
- Power supply range ±5V to ±18V
- Frequency accuracy ±1% unadjusted
- Q frequency product ≤50,000

Connection Diagrams

Ceramic Dual-In-Line Package

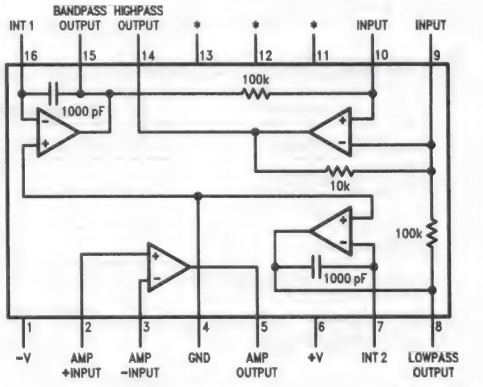


TL/K/10111–1

Top View

Order Number AF100-1CJ or AF100-2CJ
See NS Package Number HY13A

Plastic Dual-In-Line Package



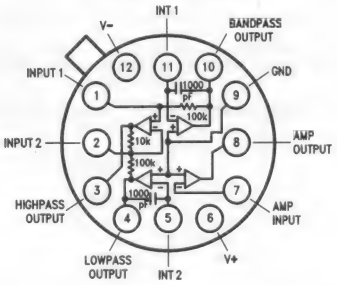
TL/K/10111–2

***Note:** Internally connected. Do not use.

Top View

Order Number AF100-1CN or AF100-2CN
See NS Package Number N16A

Metal Can Package



TL/K/10111–3

Top View

Order Number AF100-1CJ, AF100-1G, AF100-2CG or AF100-2G
See NS Package Number H12B

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	±18V
Power Dissipation	900 mW/Package (500 mW/Amp)
Differential Input Voltage	±36V
Output Short Circuit Duration (Note 1)	Infinite
Lead Temperature (Soldering, 10 sec.)	300°C

Operating Temperature	
AF100-1CJ, AF100-2CJ, AF100-1CG, AF100-2CG, AF100-1CN, AF100-2CN	−25°C to +85°C
AF100-1G, AF100-2G	−55°C to +125°C
Storage Temperature	
AF100-1G, AF100-2G	−65°C to +125°C
AF100-1CG, AF100-2CG, AF100-1CJ, AF100-2CJ, AF100-1CN, AF100-2CN	−25°C to +100°C

Electrical Characteristics (Complete Active Filter) (Note 2)

Parameter	Conditions	Min	Typ	Max	Units
Frequency Range	$f_C \times Q \leq 50{,}000$			10k	Hz
Q Range	$f_C \times Q \leq 50{,}000$			500	Hz/Hz
f_0 Accuracy AF100-1, AF100-1C AF100-2, AF100-2C	 $f_C \times Q \leq 10{,}000$, $T_A = 25°C$ $f_C \times Q \leq 10{,}000$, $T_A = 25°C$			 ±2.5 ±1.0	%
f_0 Temperature Coefficient			±50	±150	ppm/°C
Q Accuracy	$f_C \times Q \leq 10{,}000$ $T_A = 25°C$			±7.5	%
Q Temperature Coefficient			±300	±750	ppm/°C
Power Supply Current	$V_S = \pm 15V$		2.5	4.5	mA

Electrical Characteristics (Internal Op Amp) (Note 3)

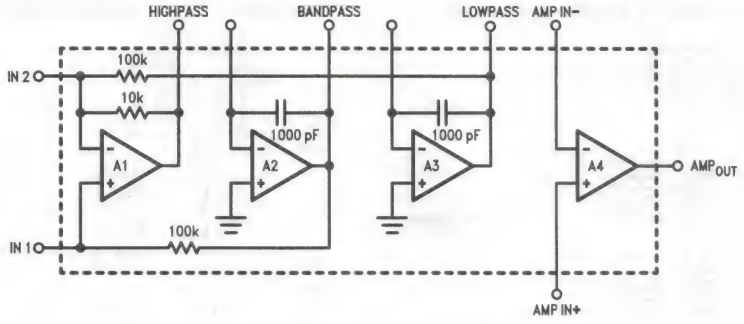
Parameter	Conditions	Min	Typ	Max	Units
Input Offset Voltage	$R_S \leq 10\ k\Omega$		1.0	6.0	mV
Input Offset Current			4	50	nA
Input Bias Current			30	200	nA
Input Resistance			2.5		MΩ
Large Signal Voltage Gain	$R_L \geq 2k$ $V_{OUT} = \pm 10V$	25	160		V/mV
Output Voltage Swing	$R_L = 10\ k\Omega$ $R_L = 2\ k\Omega$	±12 ±10	±14 ±13		V
Input Voltage Range		±12			V
Common Mode Rejection Ratio	$R_S \leq 10\ k\Omega$	70	90		dB
Supply Voltage Rejection Ratio	$R_S \leq 10\ k\Omega$	77	96		dB
Output Short Circuit Current			25		mA
Slew Rate (Unity Gain)			0.6		V/μs
Small Signal Bandwidth			1		MHz
Phase Margin			60		Degrees

Note 1: Any of the amplifiers can be shorted to ground indefinitely, however more than one should not be simultaneously shorted as the maximum junction temperature will be exceeded.

Note 2: Specifications apply for $V_S = \pm 15V$, over −25°C to +85°C for the AF100-1C and AF100-2C and over −55°C to +125°C for the AF100-1 and AF100-2, unless otherwise specified.

Note 3: Specifications apply for $V_S = \pm 15V$, $T_A = 25°C$.

Application Information



TL/K/10111–4

FIGURE 1. AF100 Schematic

CIRCUIT DESCRIPTION AND OPERATION

A schematic of the AF100 is shown in *Figure 1*. Amplifier A1 is a summing amplifier with inputs from integrator A2 to the non-inverting input and integrator A3 to the inverting input. Amplifier A4 is an uncommitted amplifier.

By adding external resistors the circuit can be used to generate the second order system

$$T(s) = \frac{a_3 s^2 + a_2 s + a_1}{s^2 + b_2 s + b_1}$$

The denominator coefficients determine the complex pole pair location and the quality of the poles where

$\omega_0 = \sqrt{b_1}$ = the radian center frequency

$Q = \frac{\omega_0}{b_2}$ = the quality of the complex pole pair

If the output is taken from the output of A1, numerator coefficients a_1 and a_2 equal zero, and the transfer function becomes:

$$T(s) = \frac{a_3 s^2}{s^2 + \frac{\omega_0}{Q} s + \omega_0^2} \quad \text{(highpass)}$$

If the output is taken from the output of A2, numerator coefficients a_1 and a_3 equal zero and the transfer function becomes:

$$T(s) = \frac{a_2 s}{s^2 + \frac{\omega_0}{Q} s + \omega_0^2} \quad \text{(bandpass)}$$

If the output is taken from the output of A3, numerator coefficients a_3 and a_2 equal zero and the transfer function becomes:

$$T(s) = \frac{a_1}{s^2 + \frac{\omega_0}{Q} s + \omega_0^2} \quad \text{(lowpass)}$$

Using proper input and output connections the circuit can also be used to generate the transfer functions for a notch and allpass filter.

In the transfer function for a notch function a_2 becomes zero, a_1 equals 1, and a_3 equals ω_Z^2. The transfer function becomes:

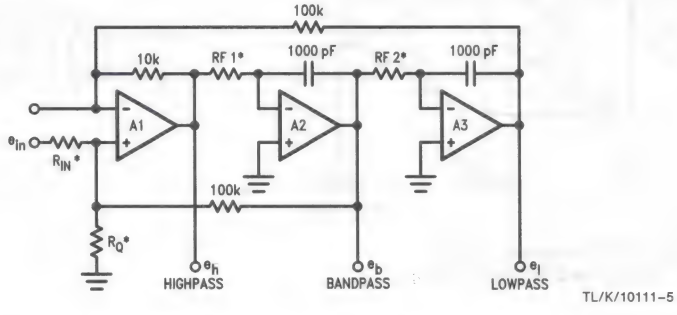
$$T(s) = \frac{s^2 + \omega_Z^2}{s^2 + \frac{\omega_0}{Q} s + \omega_0^2} \quad \text{(notch)}$$

In the allpass transfer function $a_1 = 1$, $a_2 = -\omega_0/Q$ and $a_3 = \omega_0^2$. The transfer function becomes:

$$T(s) = \frac{s^2 - \frac{\omega_0}{Q} s + \omega_0^2}{s^2 + \frac{\omega_0}{Q} s + \omega_0^2} \quad \text{(allpass)}$$

COMMON CONFIGURATIONS

The specific transfer functions for some of the most useful circuit configurations using the AF100 are illustrated in *Figures 2* through *8*. Also included are the gain equations for each transfer function in the frequency band of interest, the Q equation, center frequency equation and the Q determining resistor equation.



TL/K/10111–5

*External components

FIGURE 2. Non-Inverting Input ($Q > Q_{MIN}$, see Q Tuning Section)

Applications Information (Continued)

a) Non-inverting input *(Figure 2)* transfer equations are:

$$\frac{e_h}{e_{IN}} = \frac{s^2\left[\frac{1.1}{1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{RQ}}\right]}{\Delta} \quad \text{(highpass)}$$

$$\frac{e_b}{e_{IN}} = \frac{-s\,\omega_1\left[\frac{1.1}{1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{RQ}}\right]}{\Delta} \quad \text{(bandpass)}$$

$$\frac{e_\ell}{e_{IN}} = \omega_1\omega_2\frac{\left[\frac{1.1}{1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{RQ}}\right]}{\Delta} \quad \text{(lowpass)}$$

$$\omega_1 = \frac{10^9}{R_{F1}} \qquad \omega_2 = \frac{10^9}{R_{F2}}$$

where

$$\Delta = s^2 + s\left[\frac{1.1}{1+\frac{10^5}{RQ}+\frac{10^5}{R_{IN}}}\right]\omega_1 + 0.1\,\omega_1\omega_2$$

$$\left.\frac{e_\ell}{e_{IN}}\right|_{s\to 0} = \frac{11}{\left(1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{RQ}\right)}$$

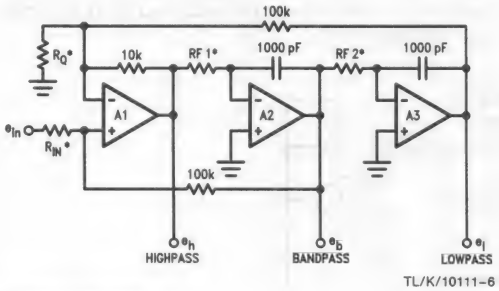
$$\left.\frac{e_h}{e_{IN}}\right|_{s\to \infty} = \frac{11}{\left(1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{RQ}\right)}$$

$$\left.\frac{e_b}{e_{IN}}\right|_{\omega=\omega_0} = -\frac{\left(1+\frac{10^5}{RQ}+\frac{10^5}{R_{IN}}\right)}{\left(1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{RQ}\right)}$$

$$\omega_0 = \sqrt{0.1\,\omega_1\omega_2}$$

$$Q = \left(\frac{1+\frac{10^5}{R_{IN}}+\frac{10^5}{RQ}}{1.1}\right)\sqrt{0.1\left(\frac{\omega_2}{\omega_1}\right)}$$

$$RQ = \frac{10^5}{\left(\frac{1.1Q}{\sqrt{0.1\frac{\omega_2}{\omega_1}}}\right) - 1 - \frac{10^5}{R_{IN}}}$$



TL/K/10111–6

*External components

FIGURE 3. Non-Inverting Input ($Q < Q_{MIN}$, see Q Tuning Section)

b) Non-inverting input *(Figure 3)* transfer equations are:

$$\frac{e_h}{e_{IN}} = \frac{s^2\left[\frac{1.1+\frac{10^4}{RQ}}{1+\frac{R_{IN}}{10^5}}\right]}{\Delta} \quad \text{(highpass)}$$

$$\frac{e_b}{e_{IN}} = \frac{-s\,\omega_1\left[\frac{1.1+\frac{10^4}{RQ}}{1+\frac{R_{IN}}{10^5}}\right]}{\Delta} \quad \text{(bandpass)}$$

$$\frac{e_\ell}{e_{IN}} = \frac{\omega_1\omega_2\left[\frac{1.1+\frac{10^4}{RQ}}{1+\frac{R_{IN}}{10^5}}\right]}{\Delta} \quad \text{(lowpass)}$$

$$\omega_1 = \frac{10^9}{R_{F1}} \qquad \omega_2 = \frac{10^9}{R_{F2}}$$

where

$$\Delta = s^2 + s\,\omega_1\left[\frac{1.1+\frac{10^4}{RQ}}{1+\frac{10^5}{R_{IN}}}\right] + 0.1\,\omega_1\omega_2$$

$$\left.\frac{e_\ell}{e_{IN}}\right|_{s\to 0} = \frac{1.1+\frac{10^4}{RQ}}{0.1\left(1+\frac{R_{IN}}{10^5}\right)}$$

$$\left.\frac{e_h}{e_{IN}}\right|_{s\to \infty} = \frac{1.1+\frac{10^4}{RQ}}{1+\frac{R_{IN}}{10^5}}$$

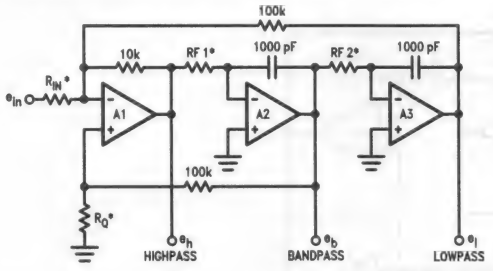
$$\left.\frac{e_b}{e_{IN}}\right|_{\omega=\omega_0} = -\frac{1+\frac{10^4}{R_{IN}}}{1+\frac{R_{IN}}{10^5}}$$

$$\omega_0 = \sqrt{0.1\,\omega_1\omega_2}$$

$$Q = \left[\frac{1+\frac{10^5}{R_{IN}}}{1.1+\frac{10^4}{RQ}}\right]\sqrt{0.1\frac{\omega_2}{\omega_1}}$$

$$RQ = \frac{10^4}{\left(1+\frac{10^5}{R_{IN}}\right)\left(\frac{\sqrt{0.1\frac{\omega_2}{\omega_1}}}{Q}\right) - 1.1}$$

Applications Information (Continued)



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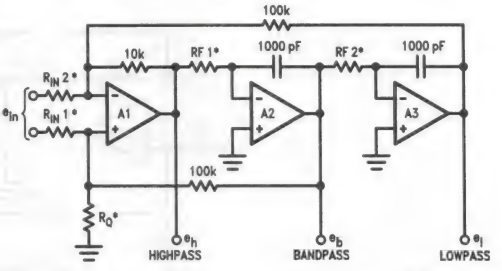
*External components

FIGURE 4. Inverting Input



TL/K/10111–8

*External components

FIGURE 5. Differential Input

c) Inverting input *(Figure 4)* transfer function equations are:

$$\frac{e_h}{e_{IN}} = \frac{-s^2 \dfrac{10^4}{R_{IN}}}{\Delta} \quad \text{(highpass)}$$

$$\frac{e_b}{e_{IN}} = \frac{s\,\omega_1 \dfrac{10^4}{R_{IN}}}{\Delta} \quad \text{(bandpass)}$$

$$\frac{e_\ell}{e_{IN}} = \frac{-\omega_1\omega_2 \dfrac{10^4}{R_{IN}}}{\Delta} \quad \text{(lowpass)}$$

$$\omega_1 = \frac{10^9}{R_{F1}} \qquad \omega_2 = \frac{10^9}{R_{F2}}$$

where

$$\Delta = s^2 + s\,\omega_1 \left[\frac{1.1 + \dfrac{10^4}{R_{IN}}}{1 + \dfrac{10^5}{R_Q}}\right] + 0.1\,\omega_1\omega_2$$

$$\left.\frac{e_\ell}{e_{IN}}\right|_{s \to 0} = -\frac{10^5}{R_{IN}} \quad \text{(lowpass)}$$

$$\left.\frac{e_h}{e_{IN}}\right|_{s \to \infty} = -\frac{10^4}{R_{IN}} \quad \text{(highpass)}$$

$$\left.\frac{e_b}{e_{IN}}\right|_{\omega = \omega_0} = \frac{\dfrac{10^4}{R_{IN}}\left(1 + \dfrac{10^5}{RQ}\right)}{1.1 + \dfrac{10^4}{R_{IN}}} \quad \text{(bandpass)}$$

$$\omega_0 = \sqrt{0.1\,\omega_1\omega_2}$$

$$Q = \left[\frac{1 + \dfrac{10^5}{RQ}}{1.1 + \dfrac{10^4}{R_{IN}}}\right]\sqrt{0.1\,\frac{\omega_2}{\omega_1}}$$

$$RQ = \frac{10^5}{\dfrac{Q}{\sqrt{0.1\,\dfrac{\omega_2}{\omega_1}}}\left(1.1 + \dfrac{10^4}{R_{IN}}\right) - 1}$$

d) Differential input *(Figure 5)* transfer function equations are:

$$\frac{e_h}{e_{IN}} = \frac{s^2 \dfrac{10^4}{R_{IN2}}}{\Delta} \quad \text{(highpass)}$$

$$\frac{e_b}{e_{IN}} = \frac{-s\,\omega_1 \dfrac{10^4}{R_{IN2}}}{\Delta} \quad \text{(bandpass)}$$

$$\frac{e_\ell}{e_{IN}} = \frac{-\omega_1\omega_2 \dfrac{10^4}{R_{IN2}}}{\Delta} \quad \text{(lowpass)}$$

$$\omega_1 = \frac{10^9}{R_{F1}} \qquad \omega_2 = \frac{10^9}{R_{F2}}$$

where

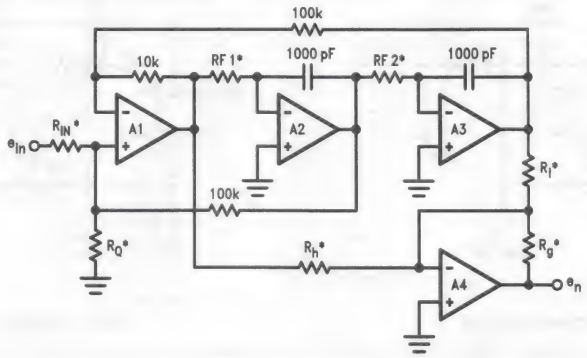
$$\Delta = s^2 + s\,\omega_1 \left[\frac{1.1 + \dfrac{10^4}{R_{IN2}}}{1 + \dfrac{10^5}{R_Q} + \dfrac{10^5}{R_{IN1}}}\right] + 0.1\,\omega_1\omega_2$$

$$\omega_0 = \sqrt{0.1\,\omega_1\omega_2}$$

$$Q = \left[\frac{1 + \dfrac{10^5}{RQ} + \dfrac{10^5}{R_{IN1}}}{1.1 + \dfrac{10^4}{R_{IN2}}}\right]\sqrt{0.1\,\frac{\omega_2}{\omega_1}}$$

$$RQ = \frac{10^5}{\dfrac{Q}{\sqrt{0.1\,\dfrac{\omega_2}{\omega_1}}}\left(1.1 + \dfrac{10^4}{R_{IN2}}\right) - 1 - \dfrac{10^5}{R_{IN1}}}$$

Applications Information (Continued)



*External components

TL/K/10111–9

FIGURE 6. Output Notch Using All Four Amplifiers

e) Output notch *(Figure 6)* transfer function equations are:

$$\frac{e_n}{e_{IN}} = \frac{\left(s^2 + \omega_Z^2\right)\left[\frac{1.1}{1 + \frac{R_{IN}}{10^5} + \frac{R_{IN}}{RQ}}\right]\frac{R_g}{R_h}}{s^2 + s\omega_1\left[\frac{1.1}{1 + \frac{10^5}{RQ} + \frac{10^5}{R_{IN}}}\right] + 0.1\,\omega_1\omega_2}$$

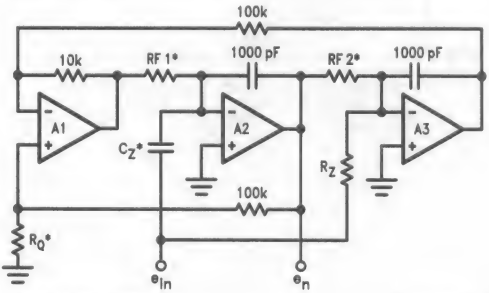
$$\omega_1 = \frac{10^9}{R_{F1}} \qquad \omega_2 = \frac{10^9}{R_{F2}} \qquad \omega_0 = \sqrt{0.1\,\omega_1\omega_2}$$

$$\omega_Z = \omega_0\sqrt{\frac{10\,R_h}{R_\ell}}$$

$$\left.\frac{e_n}{e_{IN}}\right|_{s \to 0} = \frac{11}{\left(1 + \frac{R_{IN}}{10^5} + \frac{R_{IN}}{RQ}\right)}\frac{R_g}{R_\ell}$$

$$\left.\frac{e_n}{e_{IN}}\right|_{s \to \infty} = \frac{1.1}{\left(1 + \frac{R_{IN}}{10^5} + \frac{R_{IN}}{RQ}\right)}\frac{R_g}{R_h}$$

$$\left.\frac{e_n}{e_{IN}}\right|_{\omega = \omega_Z} = 0$$



*External components

TL/K/10111–10

FIGURE 7. Input Notch Using Three Amplifiers

f) Input notch *(Figure 7)* transfer function equations are:

$$\frac{e_n}{e_{IN}} = \frac{\frac{C_Z}{10^{-9}}\left[s_2 + \omega_Z^2\right]}{s^2 + s\,\omega_1\left[\frac{1.1\,RQ}{10^5 + RQ}\right] + \omega_0^2}$$

$$\omega_1 = \frac{10^9}{R_{F1}} \qquad \omega_2 = \frac{10^9}{R_{F2}}$$

$$\omega_Z = \omega_0\sqrt{\frac{RF2 \times 10^{-9}}{R_Z\,C_Z}} \qquad \omega_0 = \sqrt{0.1\,\omega_1\omega_2}$$

$$\left.\frac{e_n}{e_{IN}}\right|_{\omega \to 0} = -\frac{R_{F2}}{R_Z}$$

$$\left.\frac{e_n}{e_{IN}}\right|_{\omega \to \infty} = -\frac{C_Z}{10^{-9}}$$

g) Allpass *(Figure 8)* transfer function equations are:

$$\frac{e_o}{e_{IN}} = -\left[\frac{s^2 - s\,\omega_1\left[\frac{1.1}{2 + \frac{R_{IN}}{RQ}}\right] + \omega_0^2}{s^2 + s\,\omega_1\left[\frac{1.1}{2 + \frac{R_{IN}}{RQ}}\right] + \omega_0^2}\right]$$

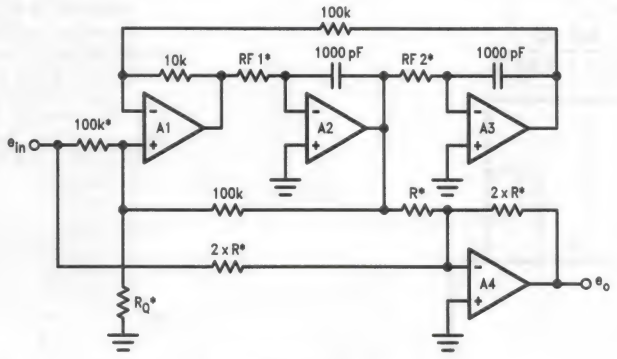
$$Q = \frac{2 + \frac{10^5}{RQ}}{1.1}\sqrt{0.1\,\frac{\omega_2}{\omega_1}}$$

$$\omega_1 = \frac{10^9}{R_{F1}} \qquad \omega_2 = \frac{10^9}{R_{F2}}$$

$$\omega_0 = \sqrt{0.1\,\omega_1\omega_2}$$

Time delay at $\omega = \frac{2Q}{\omega_0}$ seconds

Applications Information (Continued)



*External components

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FIGURE 8. Allpass

FREQUENCY TUNING

To tune the AF100 two resistors are required for frequencies between 200 Hz and 10 kHz. For lower frequencies "T" tuning or addition of external capacitors is required. Using external capacitors allows the user to go as low in frequency as he desires. "T" tuning and external capacitors can be used together.

Two resistor tuning for 200 Hz to 10 kHz

$$R_f = \frac{50.33 \times 10^6}{f_O} \Omega$$

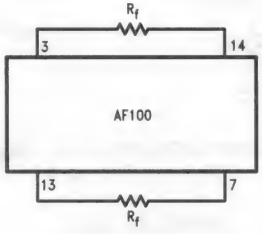

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FIGURE 9. Resistive Tuning

"T" resistive tuning for $f_O < 200$ Hz

$$R_S = \frac{R_t^2}{R_f - 2R_t} \qquad R_t < \frac{R_F}{2}$$

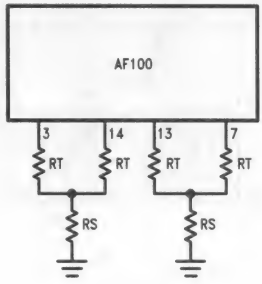

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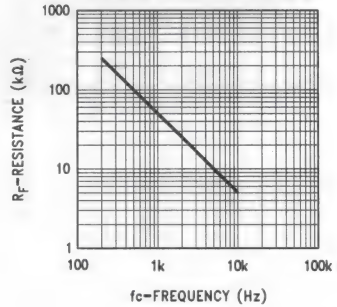
FIGURE 10. T Tuning

GRAPH A. Resistive Tuning



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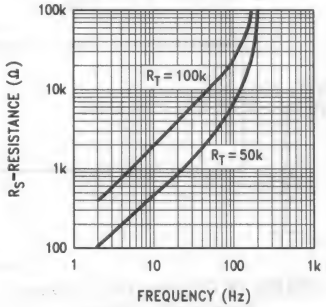
GRAPH B. "T" Tuning



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Applications Information (Continued)

RC tuning for $f_O < 200$ Hz

$$R_f = \frac{0.05033}{f_O (C + 1 \times 10^{-9})}$$

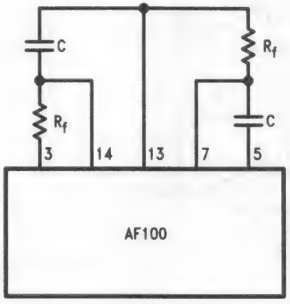

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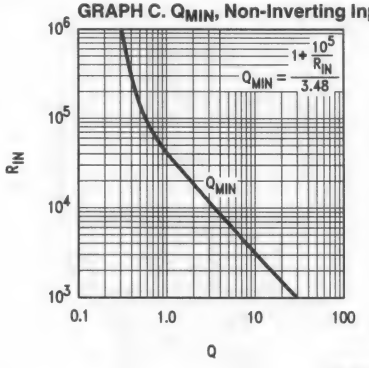
FIGURE 11. Low Frequency RC Tuning

Q TUNING

To tune the Q of an AF100 requires one resistor from pins 1 or 2 to ground. The value of the Q tuning resistor depends on the input connection and input resistance as well as the value of the Q. The Q of the unit is inversely proportional to resistance to ground at pin 1 and directly proportional to resistance to ground from pin 2.

GRAPH C. Q_{MIN}, Non-Inverting Input



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For $Q > Q_{MIN}$ in non-inverting mode:

$$RQ = \frac{10^5}{3.48Q - 1 - \frac{10^5}{R_{IN}}}$$

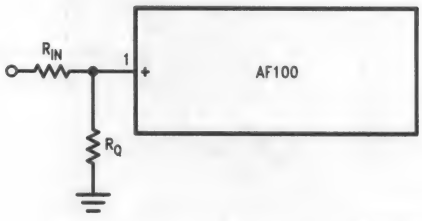

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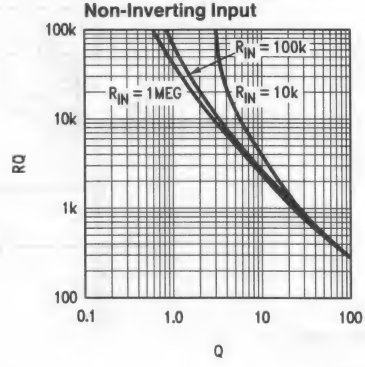
FIGURE 12. Q Tuning for $Q > Q_{MIN}$, Non-Inverting Input

GRAPH D. $Q > Q_{MIN}$, Non-Inverting Input



TL/K/10111–19

For $Q < Q_{MIN}$ in non-inverting mode:

$$RQ = \frac{10^4}{0.3162 \frac{\left(1 + \frac{10^5}{R_{IN}}\right)}{Q} - 1.1}$$

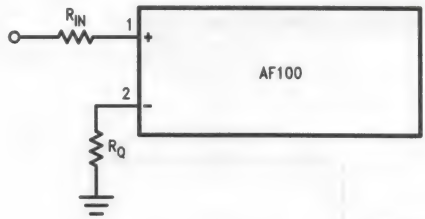


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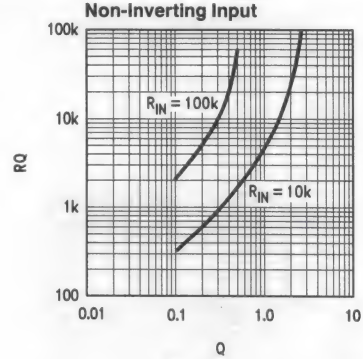
FIGURE 13. Q Tuning for $Q < Q_{MIN}$, Non-Inverting Input

GRAPH E. $Q < Q_{MIN}$, Non-Inverting Input



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Applications Information (Continued)

For any Q in inverting mode:

$$RQ = \frac{10^5}{3.16Q\left(1.1 + \frac{10^4}{R_{IN}}\right) - 1}$$

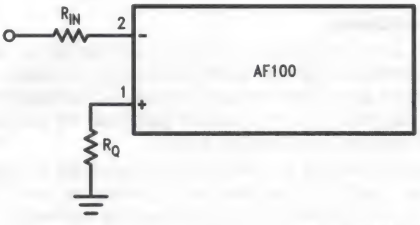

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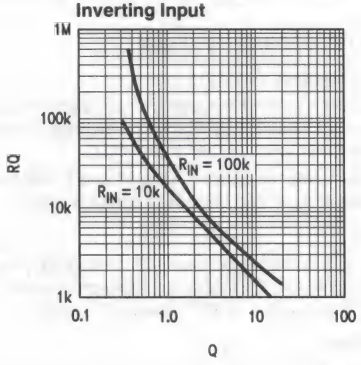
FIGURE 14. Q Tuning Inverting Input

GRAPH F. Q Tuning Inverting Input



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NOTCH TUNING

Two methods to generate notches are the RC input and lowpass/highpass summing. The RC input method requires adding a capacitor and resistor connected to the two integrator inputs. The capacitor connects to "Int 1" and the resistor connects to "Int 2". The output summing requires two resistors connected to the lowpass and highpass output.

For input RC notch tuning:

$$R_Z = \frac{R_F \times 10^{-9}}{C_Z}\left(\frac{f_O}{f_Z}\right)^2$$

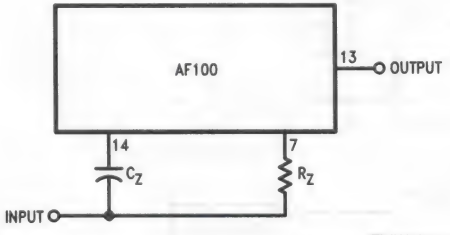

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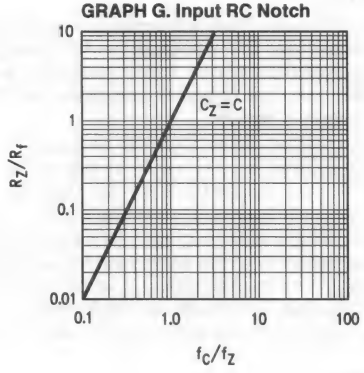
FIGURE 15. Input RC Notch

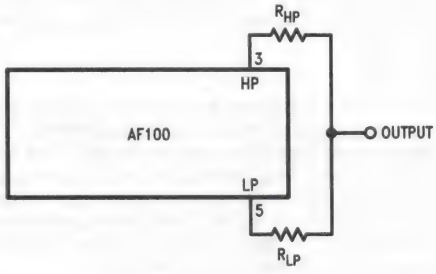
GRAPH G. Input RC Notch



TL/K/10111–25

For output notch tuning:

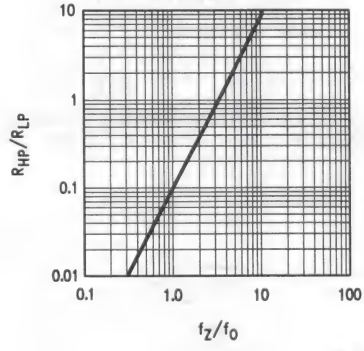
$$R_{HP} = \left(\frac{f_Z}{f_O}\right)^2 \frac{R_{LP}}{10}$$



TL/K/10111–26

FIGURE 16. Output Notch

GRAPH H. Output Notch



TL/K/10111–27

Applications Information (Continued)

TUNING TIPS

In applications where 2% to 3% accuracy is not sufficient to provide the required filter response, the AF100 stages can be tuned by adding trim pots or trim resistors in series or parallel with one of the frequency determining resistors and the Q determining resistor.

When tuning a filter section, no matter what output configuration is to be used in the circuit, measurements are made between the input and the bandpass (pin 13) output.

Before any tuning is attempted the lowpass (pin 7) output should be checked to see that the output is not clipping. At the center frequency of the section the lowpass output is 10 dB higher than the bandpass output and 20 dB higher than the highpass. This should be kept in mind because if clipping occurs the results obtained when tuning will be incorrect.

Frequency Tuning

By adjusting the resistance between pins 7 and 13 the center frequency of a section can be adjusted. If the input is through pin 1 the phase shift at center frequency will be 180° and if the input is through pin 2 the phase shift at center frequency will be 0°. Adjusting center frequency by phase is the most accurate but tuning for maximum gain is also correct.

"Q" Tuning

The "Q" is tuned by adjusting the resistance between pin 1 or 2 and ground. Low Q tuning resistors will be from pin 2 to ground ($Q < 0.6$). High Q tuning resistors will be from pin 1 to ground. To tune the Q correctly the signal source must have an output impedance very much lower than the input resistance of the filter since the input resistance affects the Q. The input must be driven through the same resistance the circuit will see to obtain precise adjustment.

The lower 3 dB (45°) frequency, f_L, and the upper 3 dB (45°) frequency, f_H, can be calculated by the following equations:

$$f_H = \left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right) \times (f_O)$$

where f_O = center frequency

$$f_L = \left(\sqrt{\left(\frac{1}{2Q}\right)^2 + 1} - \frac{1}{2Q}\right) \times (f_O)$$

When adjusting the Q, set the signal source to either f_H or f_L and adjust for 45° phase change or a 3 dB gain change.

Notch Tuning

If a circuit has a jw axis zero pair the notch can be tuned by adjusting the ratio of the summing resistors (lowpass/highpass summing) or the input resistance (input RC).

In either case the signal is connected to the input and the proper resistor is adjusted for a null at the output.

Special Cases

When using the input RC notch the unit cannot be tuned through the normal input so an additional 100k resistor can be added at pin 1 and the unit can be tuned normally. Then the 100k input resistor should be grounded and the notch tuned through the normal RC input.

An alternative way of tuning is to tune using the Q resistor as the input. This requires the Q resistor be lifted from ground and connecting the signal source to the normally grounded end of the Q resistor. This has the problem that when the Q resistor is grounded after tuning, its value is decreased by the output impedance of the source. This technique has the advantage of not requiring an additional resistor.

TUNING PROCEDURE (See *Figure 17*)

Center Frequency Tuning

Set oscillator to center frequency desired for the filter section, adjust amplitude and check that clipping does not occur at the lowpass output pin 5 (AF100J).

Adjust the resistance between pins 13 and 7 until the phase shift between input and bandpass output is 180°.

Q Tuning

Set oscillator to upper or lower 45° frequency (see tuning tips) and tune the Q resistor until the phase shift is 135° (upper 45° frequency) or 225° (lower 45° frequency).

Zero Tuning

Set the oscillator output to the zero frequency and tune the zero resistor for a null at the ouput of the summing amplifier.

Gain Adjust

Set the oscillator to any desired frequency and the gain can be adjusted by measuring the output of the summing amplifier and adjusting the feedback resistance.

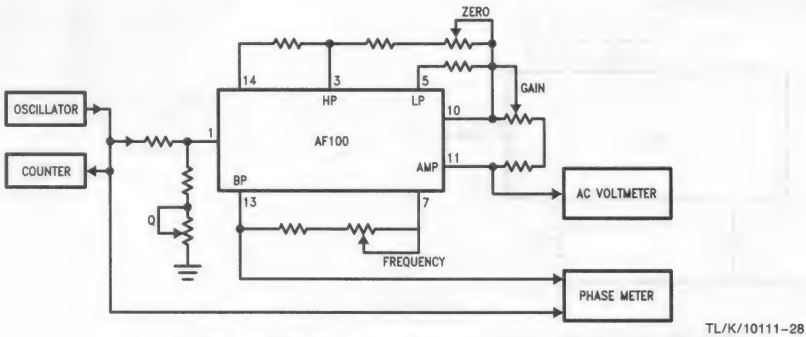


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FIGURE 17. Filter Tuning Setup

Applications Information (Continued)

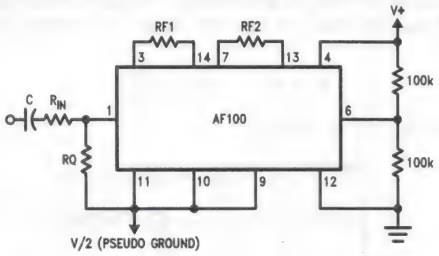

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FIGURE 18. Single Power Supply Connection Using Uncommitted Amplifier to Split Supply

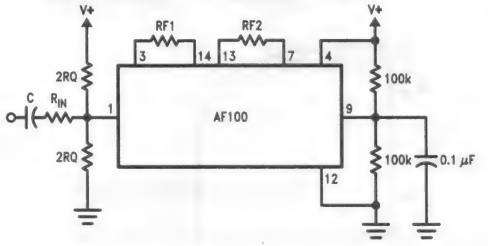

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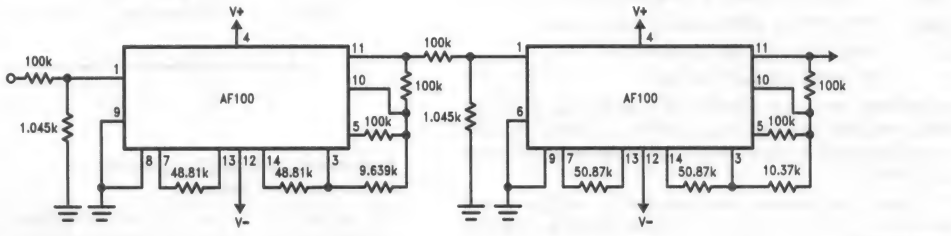
FIGURE 19. Single Power Supply Connection Using Resistive Dividers



TL/K/10111–31

Performance

0.1 dB ripple passband

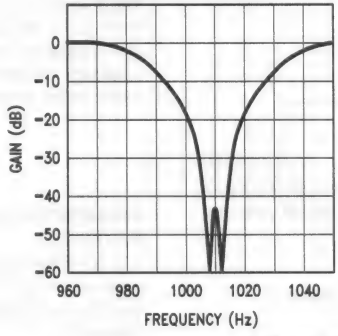
0.1 dB notch width = 100 Hz

40 dB notch width = 6.25 Hz

4th Order 1010 Hz Notch

STAGE 1

F_C = 1031.1 Hz
Q = 28.34
F_Z = 1012.2 Hz



STAGE 2

F_C = 989.3 Hz
Q = 28.34
F_Z = 1007.8 Hz

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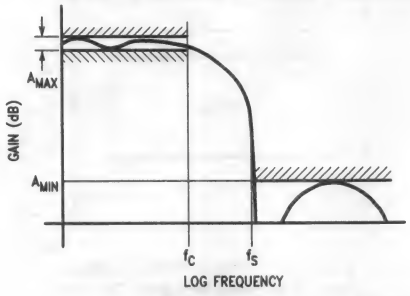
FIGURE 20. 1010 Hz Notch—Telephone Holding Tone Reject Filter

FILTER DESIGN

Since most filter tables are in terms of a normalized lowpass prototype, the filter to be designed is usually reduced to a lowpass prototype. After the lowpass transfer function is found, it is transformed to obtain the transfer function for the actual filter desired. *Graph I* shows the lowpass amplitude response which can be defined by four quantities.

Applications Information (Continued)

GRAPH I. Lowpass Prototype Response



TL/K/10111–33

A_{MAX} = the maximum peak to peak ripple in the passband.

A_{MIN} = the minimum attenuation in the stopband.

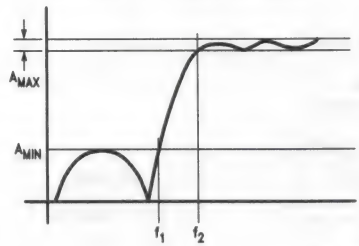
f_C = the passband cuttoff frequency.

f_S = the stopband start frequency.

By defining these four quantities for the lowpass prototype the normalized pole and zero locations and the Q (quality) of the poles can be determined from tables or by computer programs.

To obtain the lowpass prototype for the highpass filter *(Graph J)* A_{MAX} and A_{MIN} are the same as for the lowpass case but $f_C = 1/f_2$ and $f_S = 1/f_1$.

GRAPH J. Highpass Response



TL/K/10111–34

To obtain the lowpass prototype for a bandpass filter *(Graph K)* A_{MAX} and A_{MIN} are the same as for the lowpass case but

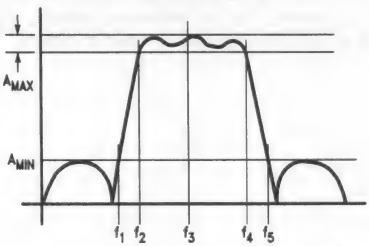
$$f_C = 1 \qquad f_S = \frac{f_5 - f_1}{f_4 - f_2}$$

where $f_3 = \sqrt{f_1 f_5} = \sqrt{f_2 f_4}$ i.e., geometric symmetry

$f_5 - f_1 = A_{MIN}$ bandwidth

$f_4 - f_2$ = Ripple bandwidth

GRAPH K. Bandpass Response



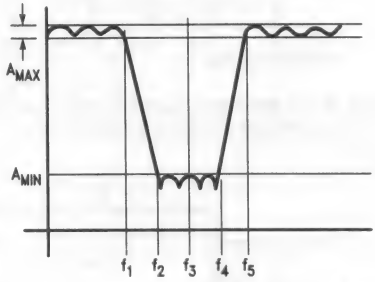
TL/K/10111–35

To obtain the lowpass prototype for the notch filter *(Graph L)* A_{MAX} and A_{MIN} are the same as for the lowpass case and

$$f_C = 1 \qquad f_S = \frac{f_5 - f_1}{f_4 - f_2}$$

where $f_3 = \sqrt{f_1 f_5} = \sqrt{f_2 f_4}$

GRAPH L. Notch Response



TL/K/10111–36

Normalized Lowpass Transformed to Un-Normalized Lowpass

The normalized lowpass filter has the passband edge normalized to unity. The un-normalized lowpass filter instead has the passband edge at f_C. The normalized and un-normalized lowpass filters are related by the transformation $s = s\omega_C$. This transforms the normalized passband edge $s = j$ to the un-normalized passband edge $s = j\omega_C$.

Normalized Lowpass Transformed to Un-Normalized Highpass

The transformation that can be used for lowpass to highpass is $S = \omega_C/s$. Since S is inversely proportional to s, the low frequency and high frequency responses are interchanged. The normalized lowpass $1/(S^2 + S/Q + 1)$ transforms to the un-normalized highpass

$$\frac{s^2}{s^2 + \frac{\omega_C}{Q}s + \omega_C{}^2}$$

Normalized Lowpass Transformed to Un-Normalized Bandpass

The transformation that can be used for lowpass to bandpass is $S = (s^2 + \omega_0{}^2)/BWs$ where $\omega_0{}^2$ is the center frequency of the desired bandpass filter and BW is the ripple bandwidth.

Normalized Lowpass Transformed to Un-Normalized Bandstop (or Notch)

The bandstop filter has a reciprocal response to a bandpass filter. Therefore a bandstop filter can be obtained by first transforming the lowpass prototype to a highpass and then performing the bandpass transformation.

SELECTION OF TRANSFER FUNCTION

The selection of a function which approximates the shape of the response desired is a complicated process. Except in the simplest cases it requires the use of tables or computer programs. The form of the transfer function desired is in terms of the pole and zero locations. The most common approximations found in tables are Butterworth, Tschebycheff, Elliptic, and Bessel. The decision as to which approximation to use is usually a function of the requirements and

Applications Information (Continued)

system objectives. Butterworth filters are the simplest but have the disadvantage of requiring high order transfer functions to obtain sharp roll-offs.

The Tschebycheff function is a min/max approximation in the passband. This approximation has the property that it is equiripple which means that the error oscillates between maximums and minimums of equal amplitude in the passband. the Tschebycheff approximation, because of its equiripple nature, has a much steeper transition region than the Butterworth approximation.

The elliptic filter, also known as Cauer or Zolotarev filters, are equiripple in the passband and stopband and have a steeper transition region than the Butterworth or the Tschebycheff.

For a specific lowpass filter three quantities can be used to determine the degree of the transfer function: the maximum passband ripple, the minimum stopband attenuation, and the transition ratio ($tr = \omega_S/\omega_C$). Decreasing A_{MAX}, increasing A_{MIN}, or decreasing tr will increase the degree of the transfer function. But for the same requirements the elliptic filter will require the lowest order transfer function. Tables and graphs are available in reference books such as "Reference Data for Radio Engineers", Howard W. Sams & Co., Inc., 5th Edition, 1970 and Erich Christian and Egon Eisenmann, "Filter Design Tables and Graphs", John Wiley and Sons, 1966.

For specific transfer functions and their pole locations such text as Louis Weinberg, "Network Analysis and Synthesis", McGraw Hill Book Company, 1962 and Richard W. Daniels, "Approximation Methods for Electronic Filter Design", McGraw-Hill Book Company, 1974, are available.

DESIGN OF CASCADED MULTISECTION FILTERS

The first step in designing is to define the response required and define the performance specifications:

1. Type of filter:
   Lowpass, highpass, bandpass, notch, allpass
2. Attenuation and frequency response
3. Performance
   Center frequency/corner frequency plus tolerance and stability
   Insertion loss/gain plus tolerance and stability
   Source impedance
   Load impedance
   Maximum output noise
   Power consumption
   Power supply voltage
   Dynamic range
   Maximum output level

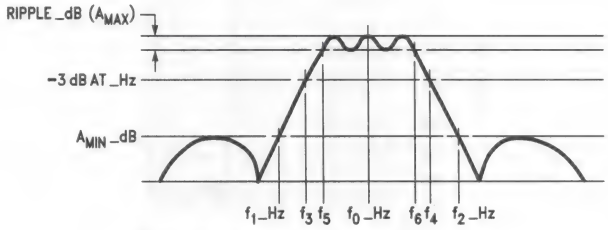
The second step is to find the pole and zero location for the transfer function which meet the above requirements. This can be done by using tables and graphs or network synthesis. The form of the transfer function which is easiest to convert to a cascaded filter is a product of first and second order terms in these forms:

First Order	Second Order	
$\frac{K}{s + \omega_R}$	$\frac{K}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$	(lowpass)
$\frac{Ks}{s + \omega_R}$	$\frac{Ks^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$	(highpass)
	$\frac{Ks}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$	(bandpass)
	$\frac{K(s^2 + \omega_Z^2)}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$	(notch)
	$\frac{s^2 - \frac{\omega_0}{Q}s + \omega_0^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$	(allpass)

Each of the second order functions is realizable by tuning an AF100 stage. By cascading these stages the desired transfer function is realized.

CASCADING SECOND ORDER STAGES

The primary concern in cascading second order stages is to minimize the maximum difference in amplitude from input to output over the frequencies of interest. A computer program is probably required in very complicated cases but some general rules that can be used that will usually give satisfactory results are:



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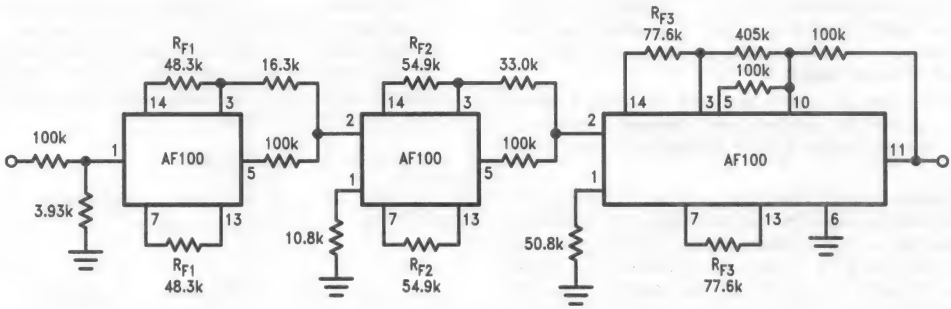
GRAPH M. Generalized Model Response

Applications Information (Continued)

1. The highest "Q" pole pair should be paired with the zero pair closest in frequency.
2. If highpass and lowpass stages are cascaded the lowpass sections should be the higher frequency and highpass sections the lower frequency.
3. In cascaded filters of more than two sections the first section should be the section with "Q" closest to 0.707 and then additional stages should be added in order of least difference between first stage Q and their Q.



TL/K/10111–38

Lowpass Elliptic Filter

$F_C = 1$
$F_S = 1.3$
$A_{MAX} = 0.1$ dB
$A_{MIN} = 40$ dB
$N = 6$

$f_{O1} = 1.0415 \quad Q_1 = 7.88 \quad f_{Z1} = 1.329 \quad f_Z/f_O = 1.28\left(\frac{f_Z}{f_O}\right)^2 = 1.63$

$f_{O2} = 0.9165 \quad Q_2 = 1.79 \quad f_{Z2} = 1.664 \quad f_Z/f_O = 1.82\left(\frac{f_Z}{f_O}\right)^2 = 3.30$

$f_{O3} = 0.649 \quad Q_3 = 0.625 \quad f_{Z3} = 4.1285 \quad f_Z/f_O = 6.36\left(\frac{f_Z}{f_O}\right)^2 = 40.5$

$$R_{F1} = \frac{(503.3)}{f_{O1} \times f_C} \times 10^5 \qquad R_{F2} = \frac{(503.3)}{f_{O2} \times f_C} \times 10^5 \qquad R_{F3} = \frac{(503.3)}{f_{O3} \times f_C}$$

at 1000 Hz = f_C

$R_{F1} = 48.3k \quad R_{F2} = 54.9k \quad R_{F3} = 77.6k$

6th Order Elliptic Filter

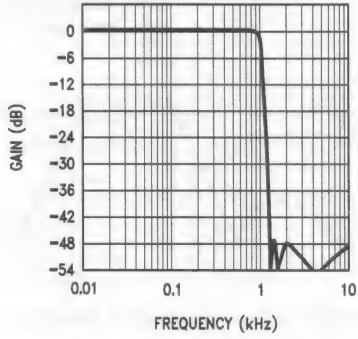


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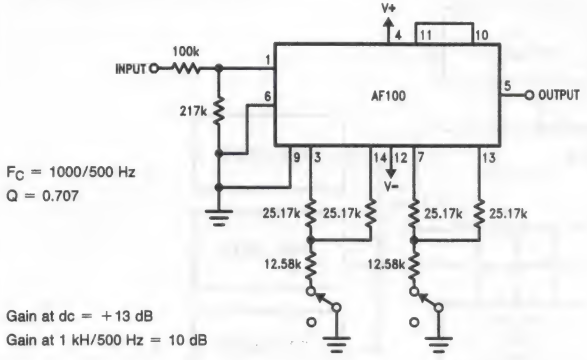
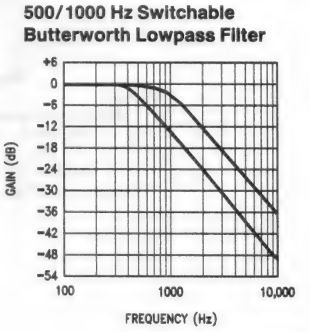
FIGURE 21. Lowpass Elliptic Filter Example

Applications Information (Continued)

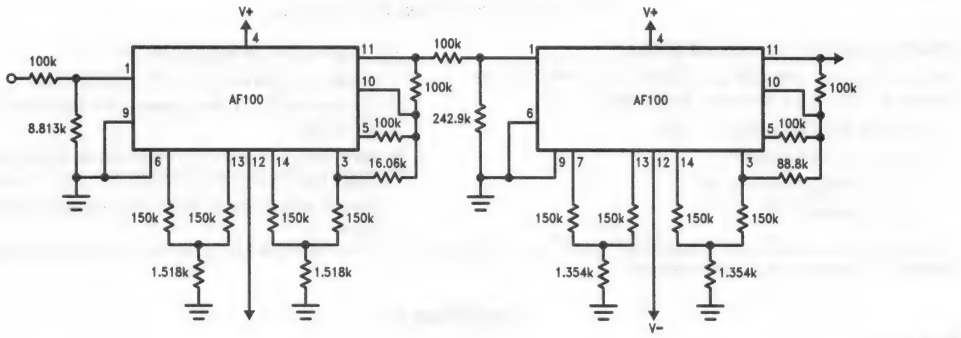




TL/K/10111–41

TL/K/10111–40

FIGURE 22. Switchable Filter Example: 500 Hz/1000 Hz Butterworth Lowpass



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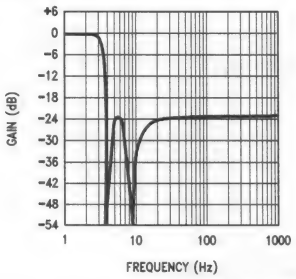


STAGE 1

F_C = 3.328 Hz
Q = 3.84
F_Z = 4.218 Hz

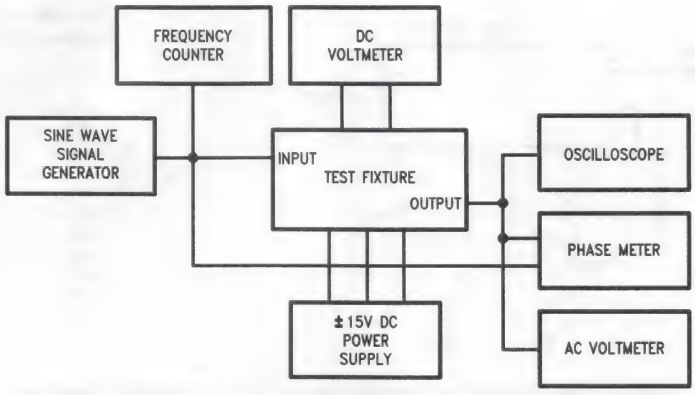
STAGE 2

F_C = 2.975 Hz
Q = 0.693
F_Z = 8.865 Hz

TL/K/10111–43

FIGURE 23. EEG Delta Filter—3 Hz Lowpass

Applications Information (Continued)



Input Level 1V rms 0 dBV

FIGURE 25. Test Circuit Block Diagram

COMPUTER AIDED DESIGN EXAMPLE*

This design is an example of a 60 Hz notch filter. The response is to have the following specifications:

Maximum passband ripple 0.1 dB

Minimum rejection 35 dB

0.1 dB bandwidth 15 Hz max

−35 dB bandwidth 1.5 Hz min

The steps in the design of this filter are:

1. Design a lowpass "prototype" for the filter.
2. Transformation of the lowpass prototype into a notch filter design.
3. Using the pole and zero locations found in step two calculate the value of the resistors required to build the filter.
4. Draw a schematic of filter using values obtained in step three.

*Computer programs shown are user interactive. Bold copy is user input, light copy is computer response, and line indications in parenthesis are included for easy identification of data common to several programs.

PROGRAM NO. 1

RUN

THIS PROGRAM DESIGNS BUTTERWORTH CHEBYCHEFF OR ELLIPTIC NORMALIZED LOWPASS FILTERS

WHAT TYPE OF FILTER? B-C-E

ELLIPTIC

DO YOU KNOW THE ORDER OF THE FILTER? Y/N

? **NO**

INPUT FC,FS,AMAX,AMIN

? **1, 10, .1, 35**

FC	1.000	
FS	10.000	
AMAX	.100	
AMIN	35.000	
N	2.000	
ATT AT FS	−35.671	(ATTENUATION IN dB)

IS THIS SATISFACTORY? Y/N

? **YES**

F Q

1.823 *(Line 1.1)* .775 *(Line 1.2)*

Z

14.124 *(Line 1.3)*

Applications Information (Continued)

PROGRAM NO. 2
(DETERMINES UN-NORMALIZED POLE + ZERO LOCATIONS OF FIRST SECTION)
(DATA ENTERED FROM PROGRAM NO. 1)

RUN
WHAT TYPE FILTER BANDPASS OR NOTCH
? **NOTCH**
ENTER # OF POLE PAIRS? **1**
ENTER # OF JW AXIS ZEROS? **1**
ENTER # OF REAL POLES? **0**
ENTER # OF ZEROS AT ZERO? **0**
ENTER # OF COMPLEX ZEROS? **0**
ENTER # OF REAL ZEROS? **0**
ENTER F&Q OF EACH POLE PAIR
? **1.823, .775** *(FROM LINE 1.1 AND LINE 1.2)*
ENTER VALUES OF JW AXIS ZEROS
? **14.124** *(FROM 1.3)*
ENTER FREQUENCY SCALING FACTOR
? **1**
ENTER THE # OF FILTERS TO BE DESIGNED
? **1**
ENTER THE C.F. AND BW OF EACH FILTER
? **60, 15**

OUTPUT OF PROGRAM NO. 2
TRANSFORMED POLE/ZERO LOCATIONS
FIRST SECTION

POLE LOCATIONS

CENTER FREQ.		Q	
56.93601	*(From Line 2.3)*	11.31813	*(From Line 2.4)*
63.228877	*(From Line 2.5)*	11.31813	*(From Line 2.6)*

JW AXIS ZEROS

59.471339	*(From Line 2.1)*
60.533361	*(From Line 2.2)*

PROGRAM NO. 3
(CHECK OF FILTER RESPONSE USING PROGRAM NO. 2 DATA BASE)

RUN
NUMERATOR [ZEROS]
A(I)S∧2+R(I)S+Z(I)∧2

1	0	59.471339	*(From Line 2.1)*
1	0	60.533361	*(From Line 2.2)*

REAL POLE

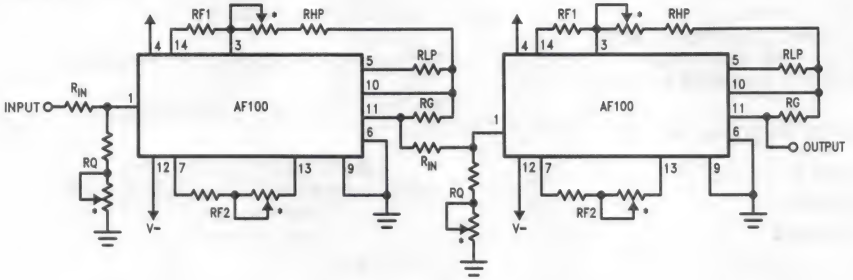
COMPLEX POLE PAIRS

	F	Q	
1	56.93601	11.31813	*(From Lines 2.3 and 2.4)*
2	63.228877	11.31813	*(From Lines 2.5 and 2.6)*

RUN

FREQ.	NOR. GAIN (DB)	PHASE	DELAY	NOR. DELAY	FREQ.	NOR. GAIN (DB)	PHASE	DELAY	NOR. DELAY
40.000	.032	347.69	.002275	5.847169	60.600	−47.102	169.17	.050801	108.232021
45.000	.060	342.20	.004107	8.749738	60.800	−33.650	165.48	.051677	110.096278
50.000	.100	330.70	.009983	21.268142	61.000	−27.577	161.72	.052809	112.508334
55.000	−.795	290.54	.046620	99.324027	61.200	−23.418	157.87	.054167	115.403169
56.000	−2.298	270.61	.063945	136.234562	61.400	−20.198	153.92	.055712	118.694436
57.000	−5.813	245.51	.072894	155.299278	61.600	−17.554	149.85	.057391	122.270086
58.000	−12.748	220.19	.065758	140.096912	61.800	−15.308	145.65	.059136	125.989157
58.200	−14.740	215.54	.063369	135.006390	62.000	−13.362	141.33	.060869	129.681062
58.400	−17.032	211.06	.060979	129.914831	63.000	−6.557	118.23	.065975	140.559984
58.600	−19.722	206.76	.058692	125.043324	64.000	−2.936	95.30	.059402	126.556312
58.800	−22.983	202.61	.056588	120.561087	65.000	−1.215	76.38	.045424	96.774832
59.000	−27.172	198.60	.054724	116.589928	66.000	−.463	62.43	.032614	69.484716
59.200	−33.235	194.72	.053139	113.212012	67.000	−.138	52.44	.023498	50.062947
59.400	−46.300	190.94	.051856	110.478482	70.000	.091	35.43	.010452	22.267368
59.600	−42.909	7.24	.050888	108.417405	75.000	.085	23.44	.004250	9.054574
59.800	−36.897	3.60	.050242	107.040235	80.000	.060	17.80	.002310	4.921727
60.00	−35.567	360.00	.049916	106.346516	85.000	.043	14.50	.001460	3.110493
60.200	−36.887	356.41	.049907	106.326777	90.000	.032	12.31	.001011	2.154297
60.400	−42.757	352.81	.050206	106.963750					

Applications Information (Continued)



*Tuning Resistors
0.1 dB bandwidth 15 Hz
−35 dB bandwidth 1.5 Hz

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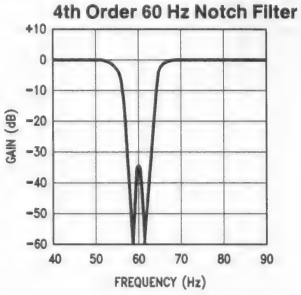
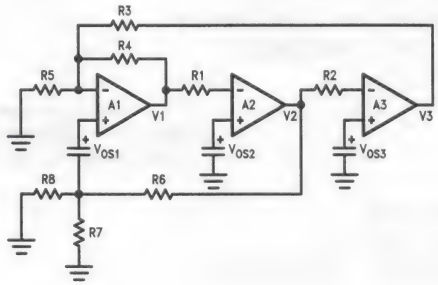


TL/K/10111–46

FIGURE 26. Implementations of a 60 Hz Notch from Computer Calculations



V1 = V_{OS2}
V2 = V_{OS3}

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$$V3 = \frac{\left[1+\left(\frac{R3+R5}{R3R5}\right)R4\right]\left[V_{OS1}+V_{OS3}\left(\frac{R7R8}{R7R8+R6(R7+R8)}\right)\right]-V_{OS2}}{\frac{R4}{R3}}$$

FIGURE 27. DC Output Voltage Due to Amplifier V_{OS}

DEFINITION OF TERMS

A_{MAX}	Maximum passband peak-to-peak ripple
A_{MIN}	Minimum stopband loss
f_Z	Frequency of jw axis pair
f_O	Frequency of complex pole pair
Q	Quality of pole
f_C	Passband edge
f_S	Stopband edge
A_{HP}	Gain from input to highpass output
A_{BP}	Gain from input to bandpass output
A_{LP}	Gain from input to lowpass output
A_{AMP}	Gain from input to output of amplifier
R_f	Pole frequency determining resistance
R_Z	Zero frequency determining resistance
R_Q	Pole quality determining resistance
f_H	Frequency above center frequency at which the gain decreases by 3 dB for a bandpass filter
f_L	Frequency below center frequency at which the gain decreases by 3 dB for a bandpass filter
BW	The bandwidth of a bandpass filter
N	Order of the denominator of a transfer function

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E. Christian and E. Eisenmann, *"Filter Design Tables and Graphs"*, John Wiley & Sons, New York, 1966

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National Semiconductor

AF150
Universal Wideband Active Filter

General Description

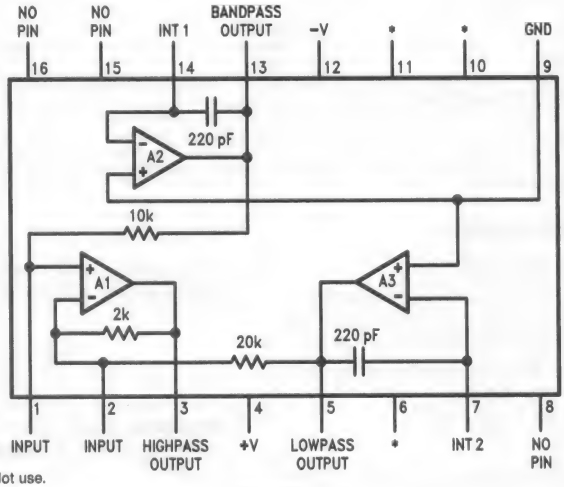
The AF150 wideband active filter is a general second order lumped RC network. Only four external resistors are required to program the AF150 for specific second order functions. Low pass, high pass and band pass functions are available simultaneously at separate outputs. Notch and all pass functions can be formed by summing the outputs with an external amplifier. Higher order filters are realized by cascading AF150 active filters with appropriate programming resistors.

Any of the classical filter configurations, such as Butterworth, Bessel, Cauer and Chebyshev can be formed.

Features

- Independent Q, frequency, gain adjustments
- Low sensitivity to external component variation
- Separate low pass, high pass, band pass outputs
- Inputs may be differential, inverting or non-inverting
- All pass and notch outputs may be formed
- Operates to 100 kHz
- Q range to 500
- Power supply range ±5V to ±18V
- High accuracy ±1% unadjusted
- Q frequency product 2×10^5

Connection Diagram



*Note: Internally connected. Do Not use.

TL/K/10112–1

Top View

Ceramic Dual-In-Line Package
Order Number AF150-1CJ or AF150-2CJ
See NS Package Number HY13A

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	±18V
Power Dissipation (Note 1)	900 mW/Package (500 mW/Amp)
Differential Input Voltage	±36V
Output Short-Circuit Duration (Note 1)	Infinite
Operating Temperature	−25°C to +85°C
Storage Temperature	−25°C to +100°C
Lead Temperature (Soldering, 10 Seconds)	300°C

Electrical Characteristics

Specifications apply for $V_S = \pm 15V$, over −25°C to +85°C unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
	Frequency Range	$f_c \times Q \leq 2 \times 10^5$			100k	Hz
	Q Range				500	Hz/Hz
	f_o Accuracy AF150-1J AF150-2J	$f_c \times Q \leq 5 \times 10^4$, $T_A = 25°C$			 ±2.5 ±1.0	%
$\Delta f_o/\Delta T$	f_o Temperature Coefficient			±50	±150	ppm/°C
	Q Accuracy	$f_c \times Q \leq 5 \times 10^4$, $T_A = 25°C$			±7.5	%
$\Delta Q/\Delta T$	Q Temperature Coefficient			±300	±750	ppm/°C
PSRR	Power Supply Rejection Ratio		80	100		dB
CMRR	Common Mode Rejection		80	100		
I_{OS}	Input Offset Current	$T_j = 25°C$		3	50	pA
I_B	Input Bias Current	$T_j = 25°C$		30	200	
V_{CM}	Input Common-Mode Voltage Range	$V_S = \pm 15V$	±11	±12		V
I_S	Power Supply Current	$V_S = \pm 15V$, $T_A = 25°C$		15	30	mA

Note 1: Any of the amplifier's outputs can be shorted to ground indefinitely; however, more than one should not be simultaneously shorted as the maximum package power dissipation will be exceeded.

Applications Information

CIRCUIT DESCRIPTION AND OPERATION

A schematic of the AF150 is shown in *Figure 1*. Amplifier A1 is a summing amplifier with inputs from integrator A2 to the non-inverting input and integrator A3 to the inverting input.

By adding external resistors the circuit can be used to generate the second order transfer function:

$$T(s) = \frac{a_3 s^2 + a_2 s + a_1}{s^2 + b_2 s + b_1}$$

The denominator coefficients determine the complex pole pair location and the quality of the poles where

$\omega_o = \sqrt{b_1}$ = the radian center frequency

$Q = \frac{\omega_o}{b_2}$ = the quality of the complex pole pair

If the output is taken from the output of A1, numerator coefficients a_1 and a_2 equal zero, and the transfer function becomes:

$$T(s) = \frac{a_3 s^2}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} \qquad \text{(high pass)}$$

If the output is taken from the output of A2, numerator coefficients a_1 and a_3 equal zero and the transfer function becomes:

$$T(s) = \frac{a_2 s}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} \qquad \text{(band pass)}$$

If the output is taken from the output of A3, numerator coefficients a_3 and a_2 equal zero and the transfer function becomes:

$$T(s) = \frac{a_1}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} \qquad \text{(low pass)}$$

Using an external op amp and the proper input and output connections, the circuit can also be used to generate the transfer functions for a notch and all pass filter.

In the transfer function for a notch function a_2 becomes zero, a_1 equals ω_z^2 and a_3 equals 1. The transfer function becomes:

$$T(s) = \frac{s^2 + \omega_z^2}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} \qquad \text{(notch)}$$

Applications Information (Continued)

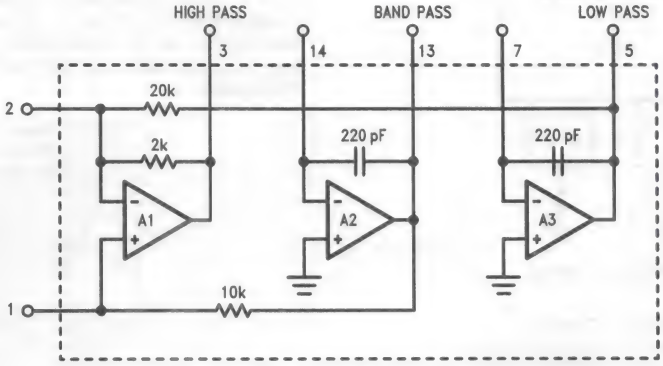


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FIGURE 1. AF150 Schematic

In the all pass transfer function $a_3 = 1$, $a_2 = -\omega_o/Q$ and $a_1 = \omega_o^2$. The transfer function becomes:

$$T(s) = \frac{s^2 - \frac{\omega_o}{Q}s + \omega_o^2}{s^2 + \frac{\omega_o}{Q}s + \omega_o^2} \quad \text{(all pass)}$$

The relationships between the generalized coefficients and the external resistors will be found in the appendix. It is not, however, necessary to use these theoretical, if not "messy", equations to solve for the proper external resistor values. In general, it is assumed that the user has knowledge of the frequency and Q of the specific filter he is designing. For higher order filters of various types, the reader is directed to any of the available texts on filters (see bibliography) for information and tables concerning the location of the poles and zeros. Once the specifics of the filter are found from the tables, it is simply a matter of cascading the sections with proper attention to some general guidelines which are included later in the application section.

The following discussion gives a step-by-step procedure for designing filters with several examples given for clarity.

FREQUENCY TUNING

Two equal value frequency setting resistors are required for frequencies above 1 kHz. For lower frequencies, T tuning or the addition of external capacitors is required. Using external capacitors allows the user to go as low in frequency as he desires. T tuning and external capacitors can be used together.

Two resistor tuning for 1 kHz to 100 kHz:

$$R_f = \frac{228.8 \times 10^6}{f_o}\ \Omega \qquad (1)$$

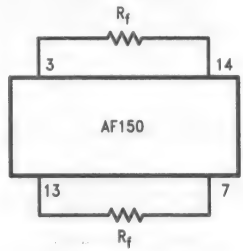

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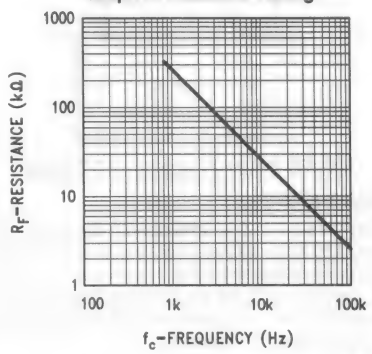
FIGURE 2. Resistive Tuning

Graph A. Resistive Tuning



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T resistive tuning for $f_o < 1$ kHz:

$$R_S = \frac{R_T^2}{R_f - 2R_T} \qquad (2)$$

R_f from equation 1.

Applications Information (Continued)

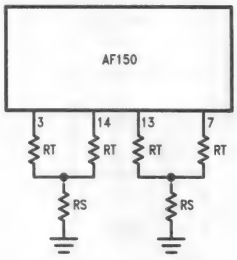

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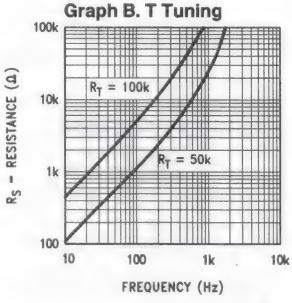
FIGURE 3. T Tuning

Graph B. T Tuning



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If external capacitors are used for $f_o < 1$ kHz, then equation 3 should be used.

$$R_f = \frac{0.05033}{f_o\,(C + 220 \times 10^{-12})}\,\Omega \qquad (3)$$

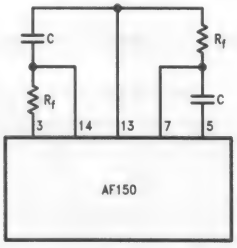

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FIGURE 4. Low Frequency RC Tuning

Q DETERMINATION

Setting the Q requires one resistor from either pin 1 or pin 2 to ground. The value of the Q setting resistor depends on the input connection and input resistance as well as the value of the Q. The Q will be inversely proportional to the resistance from pin 1 to ground and directly proportional to resistance from pin 2 to ground.

NON-INVERTING CONNECTION*

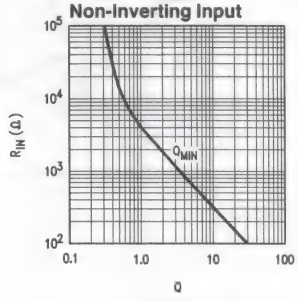
To determine the Q resistor, choose a value of input resistor, R_{IN} (*Figures 5* and *6*) and calculate Q_{MIN} (Graph C).

$$Q_{MIN} = \frac{1 + \dfrac{10^4}{R_{IN}}}{3.48}$$

***Note:** The discussion of "non-inverting" and "inverting" has to do with the phase relationship between the input port and the *low pass* output port. Refer to *Figure 1* for other output port phase relationships.

If the Q required in the circuit is greater than Q_{MIN}, use the circuit configuration shown in *Figure 5* and equation 4 to calculate R_Q, the Q resistor. If the Q of the circuit is less than Q_{MIN}, use the circuit configuration shown in *Figure 6* and equation 5.

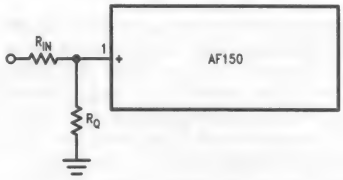
Graph C. Q_{MIN} Non-Inverting Input



TL/K/10112–8

For $Q > Q_{MIN}$ in non-inverting mode:

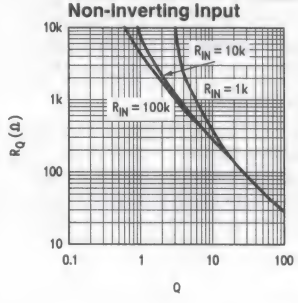
$$R_Q = \frac{10^4}{3.48Q - 1 - \dfrac{10^4}{R_{IN}}}\,\Omega \qquad (4)$$



TL/K/10112–9

FIGURE 5. Q Tuning for $Q > Q_{MIN}$, Non-Inverting Input

Graph D. R_Q for $Q > Q_{MIN}$, Non-Inverting Input



TL/K/10112–10

For $Q < Q_{MIN}$ in non-inverting mode:

$$R_Q = \frac{2 \times 10^3}{0.3162\,\dfrac{\left(1 + \dfrac{10^4}{R_{IN}}\right)}{Q} - 1.1}\,\Omega \qquad (5)$$

Applications Information (Continued)

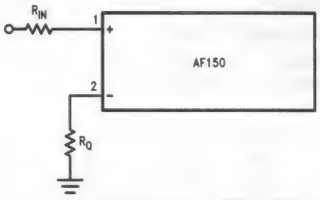


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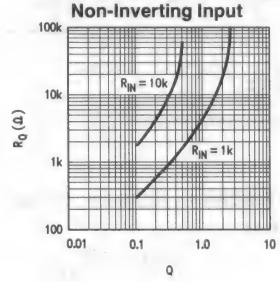
FIGURE 6. Q Tuning for Q < Q_{MIN}, Non-Inverting Input

Graph E. R_Q for Q < Q_{MIN}, Non-Inverting Input



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INVERTING CONNECTION*

For any Q in inverting mode:

$$R_Q = \frac{10^4}{3.16Q\left(1.1 + \frac{2 \times 10^3}{R_{IN}}\right) - 1}\ \Omega \qquad (6)$$

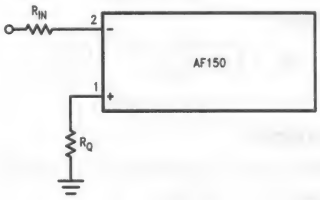

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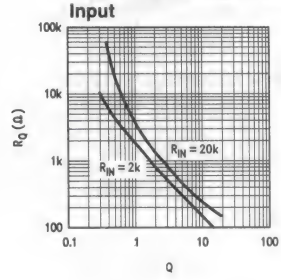
FIGURE 7. Q Tuning, Inverting Input

Graph F. Q Tuning Inverting Input



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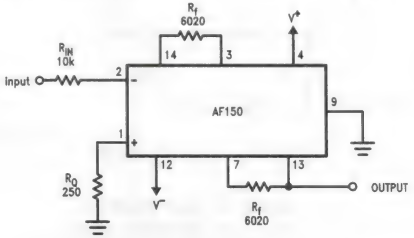
***Note:** The discussion of "non-inverting" and "inverting" has to do with the phase relationship between the input port and the *low pass* output port. Refer to *Figure 1* for other output port phase relationships.

DESIGN EXAMPLE

Non-Inverting Band Pass Filter

Center frequency 38 kHz = f_o, 10 Hz/Hz = Q, 10k = R_{IN}.



TL/K/10112–15

Using Equation 1

$$R_f = \frac{228.8 \times 10^6}{f_o}\Omega$$

$$R_f = \frac{228.8 \times 10^6}{38 \times 10^3} = 6020\Omega$$

Using Equation 6

$$R_Q = \frac{10^4}{3.16Q\left(1.1 + \frac{2 \times 10^3}{R_{IN}}\right) - 1}\ \Omega$$

$$R_Q = 250\Omega$$

From equation 33, the center frequency gain is found to be 6.3 V/V (16 dB). If the center frequency gain is to be adjusted, equation 33 can be solved for R_Q in terms of R_{IN} and this substituted into equation 6 to find the required R_{IN} and R_Q.

NOTCH FILTERS

Notches can be generated by two simple methods: using RC input *(Figure 8)* or low pass/high pass summing *(Figure 9)*. The RC input method requires adding a capacitor to pin 14 and a resistor connects to pin 7. The summing method requires two resistors connected to the low pass and high pass output.

The difference between the two possible methods of generating a notch is that the capacitor connection requires a high quality precision capacitor and the gain of the circuit is difficult to adjust because the gain and zero location are both dependent on C_Z and R_Z. The amplifier summing method requires 3 precision resistors and an external operational amplifier. However, the gain can be adjusted independent of the notch frequency.

For input RC notch tuning:

$$R_Z = \frac{C_Z R_f \times 10^{12}}{220}\left(\frac{f_o}{f_Z}\right)^2\Omega \qquad (7)$$

f_Z = frequency of notch (zero location)

Applications Information (Continued)

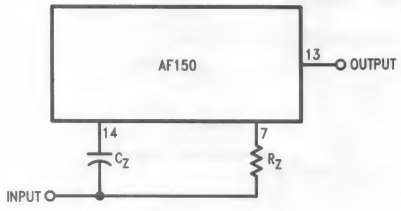


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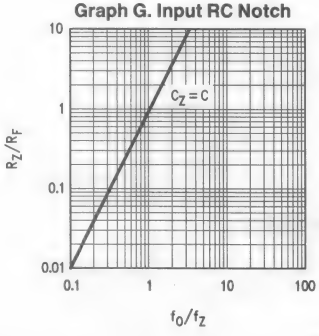
FIGURE 8. Input RC Notch

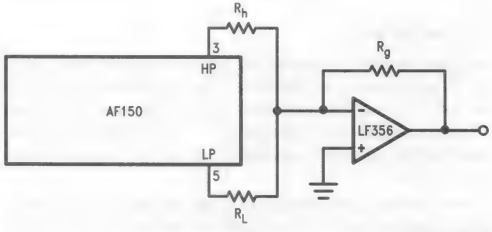
Graph G. Input RC Notch



TL/K/10112–17

For the low pass/high pass summing technique,

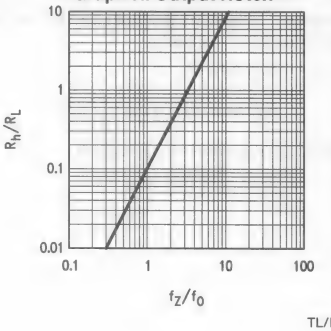
$$R_h = \left(\frac{f_Z}{f_o}\right)^2 \frac{R_L}{10} \qquad (8)$$



TL/K/10112–18

FIGURE 9. Output Notch

Graph H. Output Notch

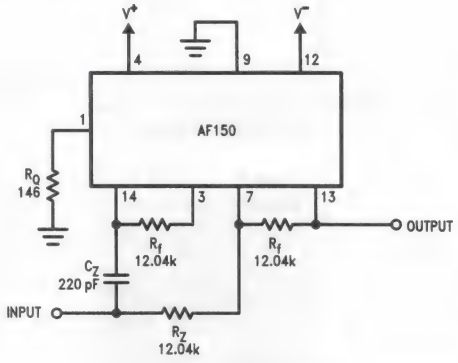


TL/K/10112–19

DESIGN EXAMPLE

19 kHz notch using RC input.

Center frequency	19 kHz	f_o
Zero frequency	19 kHz	f_Z
	20	Q



TL/K/10112–20

FIGURE 10. RC Notch, 19 kHz

Using equation 1:

$$R_f = \frac{228.8 \times 10^6}{f_o}\ \Omega$$

$$R_f = 12{,}040\Omega$$

Using equation 4 with $R_{IN} = \infty$:

$$R_Q = \frac{10^4}{3.48Q - 1 - \dfrac{10^4}{R_{IN}}}\ \Omega$$

$$R_Q = 146\Omega$$

Using equation 7:

$$R_Z = \left(\frac{C_Z R_F \times 10^{12}}{220}\right)\left(\frac{f_o}{f_Z}\right)^2\ \Omega$$

$$R_Z = 12{,}040\Omega$$

DESIGN EXAMPLE

19 kHz notch using low pass/high pass summing

Center frequency	19 kHz	f_o
Zero frequency	19 kHz	f_Z
	20	Q

Using equation 1:

$$R_f = \frac{228.8 \times 10^6}{f_o}\ \Omega$$

$$R_f = 12{,}040\Omega$$

Using equation 4, choose $R_{IN} = 10\ k\Omega$:

$$R_Q = \frac{10^4}{3.48Q - 1 - \dfrac{10^4}{R_{IN}}}\ \Omega$$

$$R_Q = 148\Omega$$

Using equation 8:

$$R_h = \left(\frac{f_Z}{f_o}\right)^2 \frac{R_L}{10}$$

Choose $R_L = 20k$, then $R_h = 2k$

Applications Information (Continued)

TRIALS, TRIBULATIONS AND TRICKS

Certainly, there is no substitute for experience when applying active filters, working with op amps or riding a bicycle. However, the following section will discuss some of the finer points in more detail, and hopefully alleviate some of the fears and problems that might be encountered.

TUNING TIPS

In applications where 2 to 3% accuracy is not sufficient to provide the required filter response, the AF150 stages can be tuned by adding trim pots or trim resistors in series or parallel with one of the frequency determining resistors and the Q determining resistor.

When tuning a filter section, no matter what output configuration is to be used in the circuit, measurements are made between the input and the band pass (pin 13) output.

Before any tuning is attempted the low pass (pin 5) output should be checked to see that the output is not clipping. At the center frequency of the section the low pass output is 10 dB higher than the band pass output and 20 dB higher than the high pass. This should be kept in mind because if clipping occurs the results obtained when tuning will be incorrect.

Frequency Tuning

By adjusting the resistance between pins 7 and 13 the center frequency of a section can be adjusted. If the input is through pin 1 the phase shift at center frequency will be 180° and if the input is through pin 2 the phase shift at center frequency will be 0°. Adjusting center frequency by phase is the most accurate but tuning for maximum gain is also correct.

Q Tuning

The Q is tuned by adjusting the resistance between pin 1 or pin 2 and ground. Low Q tuning resistors will be from pin 2 to ground ($Q < 0.6$). High Q tuning resistors will be from pin 1 to ground. To tune the Q correctly, the signal source must have an output impedance very much lower than the input resistance of the filter since the input resistance affects the Q. The input must be driven through the same resistance the circuit will see to obtain precise adjustment.

The lower 3 dB (45°) frequency, f_L, and the upper 3 dB (45°) frequency, f_H, can be calculated by the following equations:

$$f_H = \left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right) \times (f_o)$$

$$f_L = \left(\sqrt{\left(\frac{1}{2Q}\right)^2 + 1} - \frac{1}{2Q}\right) \times (f_o)$$

where f_o = center frequency

When adjusting the Q, set the signal source to either f_H or f_L and adjust for 45° phase change or a 3 dB gain change.

Notch Tuning

If a circuit has a jw axis zero pair the notch can be tuned by adjusting the ratio of the summing resistors (low pass/high pass summing) or the input resistance (input RC).

In either case the signal is connected to the input and the proper resistor is adjusted for a null at the output.

Special Cases

When using the input RC notch the unit cannot be tuned through the normal input so an additional 100k resistor can be added at pin 1 and the unit can be tuned normally. Then the 100k input resistor should be grounded and the notch tuned through the normal RC input.

An alternative way of tuning is to tune using the Q resistor as the input. This requires the Q resistor be lifted from ground and connecting the signal source to the normally grounded end of the Q resistor. This has the problem that when the Q resistor is grounded after tuning, its value is decreased by the output impedance of the source. This technique has the advantage of not requiring an additional resistor.

TUNING PROCEDURE

Center Frequency Tuning

Set oscillator to center frequency desired for the filter section, adjust amplitude and check that clipping does not occur at the low pass output pin 5.

Adjust the resistance between pins 13 and 7 until the phase shift between input and band pass output is 180°.

Q Tuning

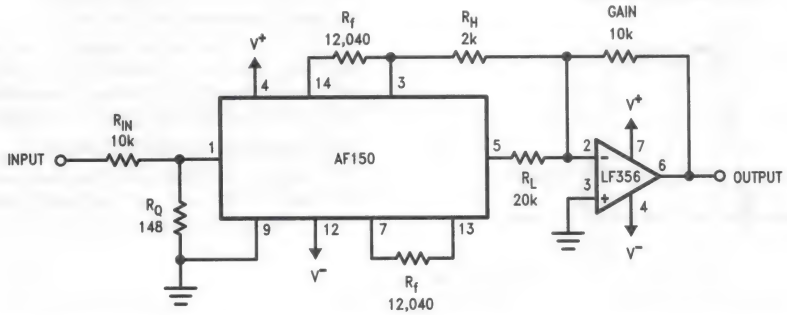
Set oscillator to upper or lower 45° frequency (see tuning tips) and tune the Q resistor until the phase shift is 135° (upper 45° frequency) or 225° (lower 45° frequency).

Zero Tuning

Set the oscillator output to the zero frequency and tune the zero resistor for a null at the output of the summing amplifier.

FILTER DESIGN

Since most filter tables are in terms of a normalized low pass prototype, the filter to be designed is usually reduced to a low pass prototype. After the low pass transfer function is found, it is transformed to obtain the transfer function for the actual filter desired. The low pass amplitude response which can be defined by four quantities, defined below:



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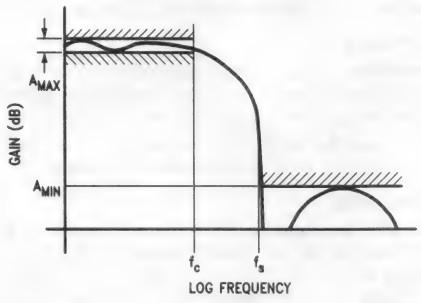
Applications Information (Continued)

Low Pass Response



TL/K/10112–22

A_{MAX} = the maximum peak-to-peak ripple in the pass band

A_{MIN} = the minimum attenuation in the stop band

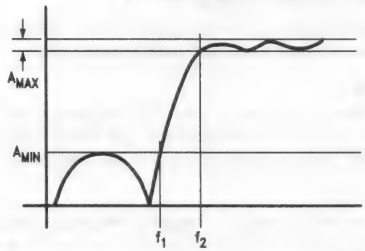
f_c = the pass band cutoff frequency

f_s = the stop band start frequency

By defining these four quantities for the low pass prototype the normalized pole and zero locations and the Q (quality) of the poles can be determined from tables or by computer programs.

To obtain the high pass from the low pass filter tables, A_{MAX} and A_{MIN} are the same as for the low pass case, but $f_c = 1/f_2$ and $f_s = 1/f_1$.

High Pass Response



TL/K/10112–23

To obtain the band pass from the low pass filter tables, A_{MAX} and A_{MIN} are the same as for the low pass case, but:

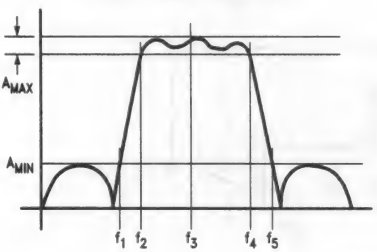
$$f_c = 1 \qquad f_s = \frac{f_5 - f_1}{f_4 - f_2}$$

where $f_3 = \sqrt{f_1 \times f_5} = \sqrt{f_2 \times f_4}$ i.e., geometric symmetry

$f_5 - f_1 = A_{MIN}$ bandwidth

$f_4 - f_2$ = Ripple bandwidth

Band Pass Response



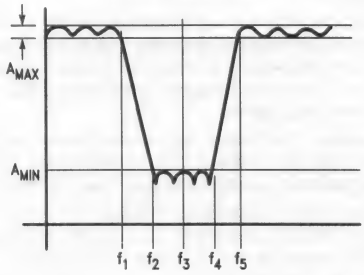
TL/K/10112–24

To obtain the notch from the low pass filter tables, A_{MAX} and A_{MIN} are the same as for the low pass case and

$$f_c = 1, \qquad f_s = \frac{f_5 - f_1}{f_4 - f_2}$$

where $f_3 = \sqrt{f_1 \times f_5} = \sqrt{f_2 \times f_4}$

Notch Response



TL/K/10112–25

Normalized Low Pass Transformed to Un-Normalized Low Pass

The normalized low pass filter has the pass band edge normalized to unity. The un-normalized low pass filter instead has the pass band edge at f_c. The normalized and un-normalized low pass filters are related by the transformation $s = s\omega_c$. This transforms the normalized pass band edge $s = j$ to the un-normalized pass band edges $s = j\omega_c$.

Normalized Low Pass Transformed to Un-Normalized High Pass

The transformation that can be used for low pass to high pass is $S = \omega_C/s$. Since S is inversely proportional to s, the low frequency and high frequency responses are interchanged. The normalized low pass $1/(S^2 + S/Q + 1)$ transforms to the un-normalized high pass.

$$\frac{s^2}{s^2 + \frac{\omega_C}{Q}S + \omega_C{}^2}$$

Normalized Low Pass Transformed to Un-Normalized Band Pass

The transformation that can be used for low pass to band pass is:

$$S = \frac{s^2 + \omega_0{}^2}{BS \times s}$$

where $\omega_0{}^2$ is the center frequency of the desired band pass filter and BW is the ripple bandwidth.

Normalized Low Pass Transformed to Un-Normalized Band Stop (Or Notch)

The bandstop filter has a reciprocal response to a band pass filter. Therefore, a bandstop filter can be obtained by first transforming the low pass prototype to a high pass and then performing the band pass transformation.

SELECTION OF TRANSFER FUNCTION

The selection of a function which approximates the shape of the response desired is a complicated process. Except in the simplest case, it requires the use of tables or computer

Applications Information (Continued)

programs. The form of the transfer function desired is in terms of the pole and zero locations. The most common approximations found in tables are Butterworth, Chebychev, Elliptic and Bessel. The decision as to which approximation to use is usually a function of the requirements and system objectives. Butterworth filters are the simplest but have the disadvantage of requiring high order transfer functions to obtain sharp roll-offs.

The Chebychev function is a min/max approximation in the pass band. This approximation has the property that it is equiripple which means that the error oscillates between maximums and minimums of equal amplitude in the pass band. The Chebychev approximation, because of its equiripple nature, has a much steeper transition region than the Butterworth approximation.

The elliptic filter, also known as Cauer or Zolotarev filters, are equiripple in the pass band and stop band and have a steeper transition region than the Butterworth or the Chebychev.

For a specific low pass filter three quantities can be used to determine the degree of the transfer function: the maximum pass band ripple, the minimum stop band attenuation, and the transition ratio ($tr = \omega_s/\omega_c$). Decreasing A_{MAX}, increasing A_{MIN}, or decreasing t_r will increase the degree of the transfer function. But for the same requirements the elliptic filter will require the lowest order transfer function. Tables and graphs are available in reference books such as "Reference Data for Radio Engineers", Howard W. Sams & Co., Inc., 5th Edition, 1970 and Erich Christian and Egon Eisenmann, "Filter Design Tables and Graphs"; John Wiley and Sons, 1966.

For specific transfer functions and their pole locations such texts as Louis Weinberg, "Network Analysis and Synthesis", McGraw Hill Book Company, 1962 and Richard W. Daniels, "Approximation Methods for Electronic Filter Design", McGraw-Hill Book Company, 1974, are available.

DESIGN OF CASCADED MULTISECTION FILTERS

The first step in designing is to define the response required and define the performance specifications:

1. Type of filter:
   Low pass, high pass, band pass, notch, all pass
2. Attenuation and frequency response
3. Performance
   Center frequency/corner frequency plus tolerance and stability
   Insertion loss/gain plus tolerance and stability
   Source impedance
   Load impedance
   Maximum output noise
   Power consumption
   Power supply voltage
   Dynamic range
   Maximum output level

The second step is to find the pole and zero location for the transfer function which meet the above requirements. This can be done by using tables and graphs or network synthesis. The form of the transfer function which is easiest to convert to a cascaded filter is a product of first and second order terms in these forms:

First Order	Second Order	
$\frac{K}{s+\omega_r}$	$\frac{K}{s^2+\frac{\omega_o}{Q}s+\omega_o^2}$	(low pass)
$\frac{K_S}{s+\omega_r}$	$\frac{K_s^2}{s^2+\frac{\omega_O}{Q}s+\omega_o^2}$	(high pass)
	$\frac{K_s}{s^2+\frac{\omega_o}{Q}s+\omega_o^2}$	(band pass)
	$\frac{K(s^2+\omega_Z^2)}{s^2+\frac{\omega_o}{Q}s+\omega_o^2}$	(notch)
	$\frac{s^2-\frac{\omega_o}{Q}s+\omega_o^2}{s^2+\frac{\omega_o}{Q}s+\omega_o^2}$	(all pass)

Each of the second order functions is realizable by using an AF150 stage. By cascading these stages the desired transfer function is realized.

CASCADING SECOND ORDER STAGES

The primary concern in cascading second order stages is to minimize the difference in amplitude from input to output over the frequencies of interest. A computer program is probably required in very complicated cases but some general rules that can be used that will usually give satisfactory results are:

1. The highest Q pole pair should be paired with the zero pair closest in frequency.
2. If high pass and low pass stages are cascaded, the low pass sections should be the higher frequency and high pass sections the lower frequency.
3. In cascaded filters of more than two sections, the first section should be the section with Q closest to 0.707 and then additional stages should be added in order of least difference between first stage Q and their Q.

DESIGN EXAMPLES OF CASCADE CONNECTIONS

Example 1:

Consider a 4th order Butterworth low pass filter with a 10 kHz cutoff (−3 dB) frequency and input impedance ≥30 kΩ.

From tables, the normalized filter parameters are:

F1 = 1.0 Q1 = 0.541

F2 = 1.0 Q2 = 1.306

Thus, relative to the design required

F1 = (1.0) (10 kHz) = 10 kHz

F2 = (1.0) (10 kHz) = 10 kHz

Section 1

$$F = 10\ \text{kHz}, \quad Q = 1.306$$

$$R_f = \frac{228.8 \times 10^6}{f_o}\ \Omega \qquad \text{(Using equation 1)}$$

$$R_f = 22{,}880\Omega$$

Applications Information (Continued)

Select input resistor 31.6 kΩ

$$Q_{MIN} = \frac{1 + \frac{10^4}{R_{IN}}}{3.48}$$

$$Q_{MIN} = 0.378$$

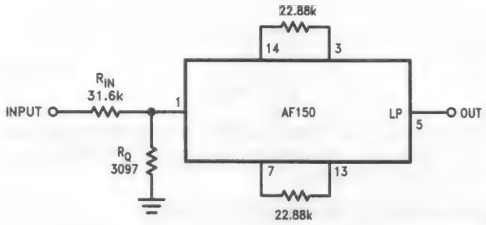
Thus, $Q > Q_{MIN}$

Therefore:

$$R_Q = \frac{10^4}{3.48Q - 1 - \frac{10^4}{R_{IN}}}\ \Omega \quad \text{(Using equation 4)}$$

$$R_Q = 3097\Omega$$

First Stage



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Section 2

$$f_o = 10k, \quad Q = 0.541$$

Since f_o is the same as for the first section:

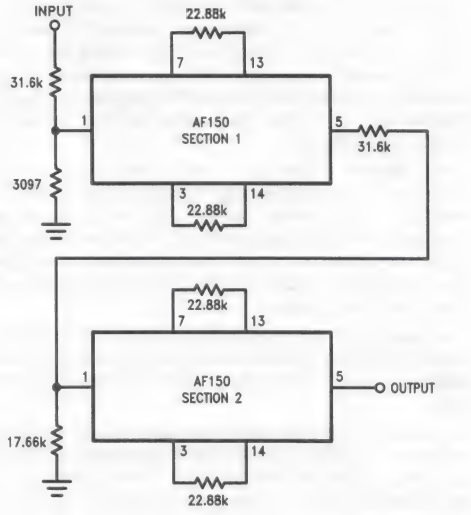
$$R_f = 22.88\ k\Omega$$

Select $R_{IN} = 31.6\ k\Omega$

$$R_Q = \frac{10^4}{3.48Q - 1 - \frac{10^4}{R_{IN}}}\ \Omega \quad \text{(Using equation 4)}$$

$$R_Q = 17{,}661\Omega$$

Complete Filter, Example 1



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Example 2.

Consider the design of a low pass filter with the following performance:

$$f_c = 10\ kHz$$

$$f_s = 11\ kHz$$

$$A_{MAX} = 1\ dB$$

$$A_{MIN} = 40\ dB$$

It is found that a 6th order elliptic filter will satisfy the above requirements. The parameters of the design are:

STAGE	f_o (kHz)	Q	f_z (kHz)
1	5.16	0.82	29.71
2	8.83	3.72	13.09
3	10.0	20.89	11.15

Applications Information (Continued)

Stage 1

a) From equation 1, R_F is found to be 44.34k

b) From equation 4, R_Q is found to be 11.72k assuming R_{IN} (arbitrary) is 10 kΩ.

To create the transmission zero f_z, at 29.71 kHz, use equation 8.

$$R_h = \left(\frac{f_z}{f_o}\right)^2 \frac{R_L}{10}, \quad \text{or } R_h = \left(\frac{29.71}{5.16}\right)^2 \frac{R_L}{10}$$

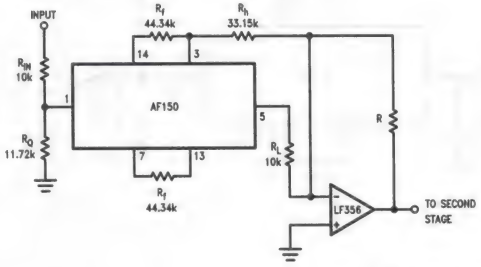
Thus,

$$R_h = 3.315\, R_L$$

If R_L is arbitrarily chosen as 10 kΩ, R_h = 33.15k.

Thus, the design of the first stage is:

First Stage



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where the feedback resistor, R, around the external op amp may be used to adjust the gain.

Stage 2

The second stage design follows exactly the same procedure as the first stage design. The results are:

a) From equation 1, R_f = 25.91k

b) From equation 4, R_Q = 913.6Ω, again assuming R_{IN} is arbitrarily 10k.

c) $R_h = \left(\frac{13.09}{8.83}\right)^2 \frac{R_L}{10}$ or $R_h = 0.22\, R_L$

Selecting R_L = 10k, then R_h = 2.2k, the second stage design is shown below.

Stage 3

The third stage design, again, is identical to the first 2 stages and the results are (for R_{IN} = 10k):

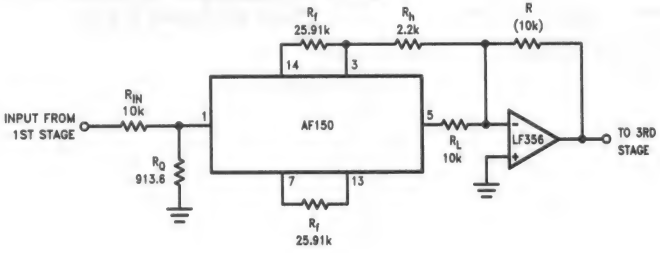
$$R_f = \frac{228.8 \times 10^6}{f_o} = 22.88k$$

$$R_Q = \frac{10^4}{3.48Q - 1 - \frac{10^4}{R_{IN}}} = 141.4\Omega$$

$$R_h = \left(\frac{f_z}{f_o}\right)^2 \frac{R_L}{10} = \left(\frac{11.5}{10}\right)^2 \frac{R_L}{10} \qquad R_h = 0.124\, R_L$$

Let R_L = 20k, R_h = 2.48k

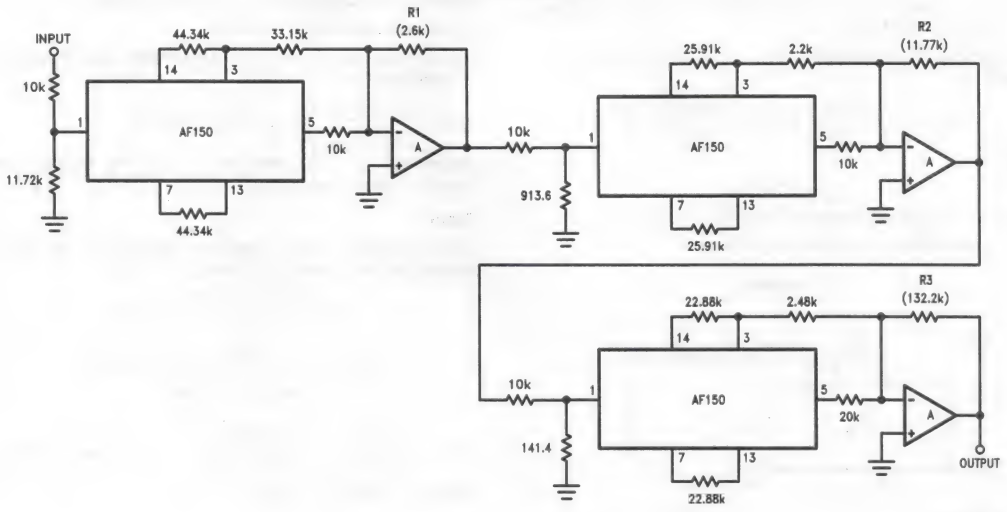
Second Stage



TL/K/10112–29

Applications Information (Continued)

Filter for Example 2



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Note 1: Select R1, R2, R3 for desired gain.

Note 2: All amplifiers LF356.

From equation 13, the DC gain of the first section is

$$A_{V1} = \frac{11}{1 + \frac{R_{IN}}{10^4} + \frac{R_{IN}}{R_Q}}$$

$$A_{V1} = \frac{11}{1 + \frac{10^4}{10^4} + \frac{10^4}{11.72 \times 10^3}} = 3.86\ V/V$$

Similarly, the DC gain of the second and third sections are:

$$A_{V2} = 0.850$$

$$A_{V3} = 0.151$$

Therefore, the overall DC gain is 0.495 and can be adjusted by selecting R1 with respect to 10k, R2 with respect to 10k or R3 with respect to 20k.

Applications Information (Continued)

For convenience, a standard resistor value table is given below.

Standard Resistance Values are obtained from the Decade Table by multiplying by multiples of 10. As an example, 1.33 can represent 1.33Ω, 133Ω, 1.33 kΩ, 13.3 kΩ, 133 kΩ, 1.33 MΩ.

Standard 5% and 2% Resistance Values

Ohms	Ohms	Ohms	Ohms	Ohms	Ohms	Ohms	Ohms	Ohms	Ohms	Ohms	Megohms	
10	27	68	180	470	1,200	3,300	8,200	22,000	56,000	150,000	0.24	0.62
11	30	75	200	510	1,300	3,600	9,100	24,000	62,000	160,000	0.27	0.68
12	33	82	220	560	1,500	3,900	10,000	27,000	68,000	180,000	0.30	0.75
13	36	91	240	620	1,600	4,300	11,000	30,000	75,000	200,000	0.33	0.82
15	39	100	270	680	1,800	4,700	12,000	33,000	82,000	220,000	0.36	0.91
16	43	110	300	750	2,000	5,100	13,000	36,000	91,000		0.39	1.0
18	47	120	330	820	2,200	5,600	15,000	39,000	100,000		0.43	1.1
20	51	130	360	910	2,400	6,200	16,000	43,000	110,000		0.47	1.2
22	56	150	390	1,000	2,700	6,800	18,000	47,000	120,000		0.51	1.3
24	62	160	430	1,100	3,000	7,500	20,000	51,000	130,000		0.56	1.5

Decade Table Determining ½% and 1% Standard Resistance Values

1.00	1.21	1.47	1.78	2.15	2.61	3.16	3.83	4.64	5.62	6.81	8.25
1.02	1.24	1.50	1.82	2.21	2.67	3.24	3.92	4.75	5.76	6.98	8.45
1.05	1.27	1.54	1.87	2.26	2.74	3.32	4.02	4.87	5.90	7.15	8.66
1.07	1.30	1.58	1.91	2.32	2.80	3.40	4.12	4.99	6.04	7.32	8.87
1.10	1.33	1.62	1.96	2.37	2.87	3.48	4.22	5.11	6.19	7.50	9.09
1.13	1.37	1.65	2.00	2.43	2.94	3.57	4.32	5.23	6.34	7.68	9.31
1.15	1.40	1.69	2.05	2.49	3.01	3.65	4.42	5.36	6.49	7.87	9.53
1.18	1.43	1.74	2.10	2.55	3.09	3.74	4.53	5.49	6.65	8.06	9.76

Appendix (See Footnote)

The specific transfer functions for some of the most useful circuit configurations using the AF150 are illustrated in *Figures 11–17*. Also included are the gain equations for each transfer function in the frequency band of interest, the Q equation, center frequency equation and the Q determining resistor equation. Q_{MIN} is a function of R_{IN} (see Graph C).

a. Non-Inverting Input *(Figure 10)*
Transfer Equations are:

$$\frac{e_h}{e_{IN}} = \frac{s^2\left[\dfrac{1.1}{1+\dfrac{R_{IN}}{10^4}+\dfrac{R_{IN}}{R_Q}}\right]}{\Delta} \text{ (high pass)} \quad (9)$$

$$\frac{e_b}{e_{IN}} = \frac{-s\,\omega_1\left[\dfrac{1.1}{1+\dfrac{R_{IN}}{10^4}+\dfrac{R_{IN}}{R_Q}}\right]}{\Delta} \text{ (band pass)} \quad (10)$$

$$\frac{e_\ell}{e_{IN}} = \frac{\omega_1\,\omega_2\left[\dfrac{1.1}{1+\dfrac{R_{IN}}{10^4}+\dfrac{R_{IN}}{R_Q}}\right]}{\Delta} \text{ (low pass)} \quad (11)$$

where

$$\Delta = s^2 + s\left[\frac{1.1}{1+\frac{10^4}{R_Q}+\frac{10^4}{R_{IN}}}\right]\omega_1 + 0.1\,\omega_1\,\omega_2 \quad (12)$$

$$\left.\frac{e_\ell}{e_{IN}}\right|_{s\to 0} = \frac{11}{\left(1+\dfrac{R_{IN}}{10^4}+\dfrac{R_{IN}}{R_Q}\right)} \text{ (DC Gain)} \quad (13)$$

$$\left.\frac{e_h}{e_{IN}}\right|_{s\to \infty} = \frac{1.1}{\left(1+\dfrac{R_{IN}}{10^4}+\dfrac{R_{IN}}{R_Q}\right)} \text{ (High Freq. Gain)} \quad (14)$$

$$\left.\frac{e_b}{e_{IN}}\right|_{\omega=\omega_0} = -\frac{\left(1+\dfrac{10^4}{R_Q}+\dfrac{10^4}{R_{IN}}\right)}{\left(1+\dfrac{R_{IN}}{10^4}+\dfrac{R_{IN}}{R_Q}\right)} \text{ (Center Freq. Gain)} \quad (15)$$

$$\omega_1 = \frac{10^{12}}{R_{f1}\times 220} \qquad \omega_2 = \frac{10^{12}}{R_{f2}\times 220}$$

where

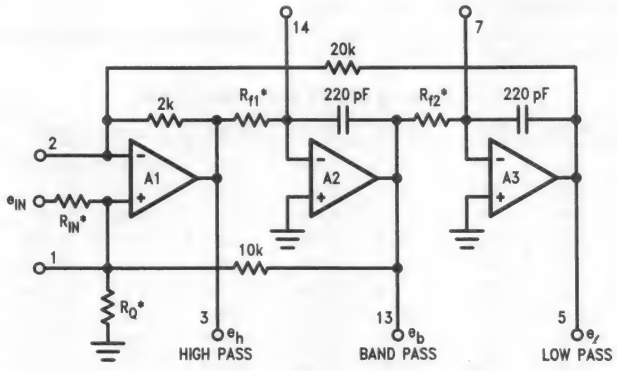
$\omega_0 = \sqrt{0.1\,\omega_1\,\omega_2}$, (see Footnote)

$$Q = \left(\frac{1+\dfrac{10^4}{R_{IN}}+\dfrac{10^4}{R_Q}}{1.1}\right)\sqrt{0.1\left(\frac{\omega_2}{\omega_1}\right)} \quad (16)$$

$$R_Q = \frac{10^4}{\left(\dfrac{1.1Q}{\sqrt{0.1\dfrac{\omega_2}{\omega_1}}}\right) - 1 - \dfrac{10^4}{R_{IN}}} \quad (17)$$

Note: It should be noted that in the text of this paper, ω_1 and ω_2 have been assumed equal, and hence $R_{f1} = R_{f2}$. No generality is lost in this assumption and it facilitates the design. However, for completeness, the equations given are exact.

Appendix (Continued)



*External Components

FIGURE 11. Non-Inverting Input ($Q > Q_{MIN}$)

b) Non-inverting input *(Figure 12)* transfer equations are:

$$\frac{e_h}{e_{IN}} = \frac{s^2 \left[\frac{1.1 + \frac{2 \times 10^3}{R_Q}}{1 + \frac{R_{IN}}{10^4}} \right]}{\Delta} \text{ (high pass)} \quad (18)$$

$$\frac{e_b}{e_{IN}} = \frac{-s\,\omega_1 \left[\frac{1.1 + \frac{2 \times 10^3}{R_Q}}{1 + \frac{R_{IN}}{10^4}} \right]}{\Delta} \text{ (band pass)} \quad (19)$$

$$\frac{e_\ell}{e_{IN}} = \frac{\omega_1 \omega_2 \left[\frac{1.1 + \frac{2 \times 10^3}{R_Q}}{1 + \frac{R_{IN}}{10^4}} \right]}{\Delta} \text{ (low pass)} \quad (20)$$

where

$$\Delta = s^2 + s\,\omega_1 \left[\frac{1.1 + \frac{2 \times 10^3}{R_Q}}{1 + \frac{10^4}{R_{IN}}} \right] + 0.1\,\omega_1 \omega_2 \quad (21)$$

$$\left.\frac{e_\ell}{e_{IN}}\right|_{s \to 0} = \frac{1.1 + \frac{2 \times 10^3}{R_Q}}{0.1\left(1 + \frac{R_{IN}}{10^4}\right)} \quad (22)$$

$$\left.\frac{e_h}{e_{IN}}\right|_{s \to \infty} = \frac{1.1 + \frac{2 \times 10^3}{R_Q}}{1 + \frac{R_{IN}}{10^4}} \quad (23)$$

$$\left.\frac{e_b}{e_{IN}}\right|_{\omega = \omega_0} = -\frac{1 + \frac{10^4}{R_{IN}}}{1 + \frac{R_{IN}}{10^4}} \quad (24)$$

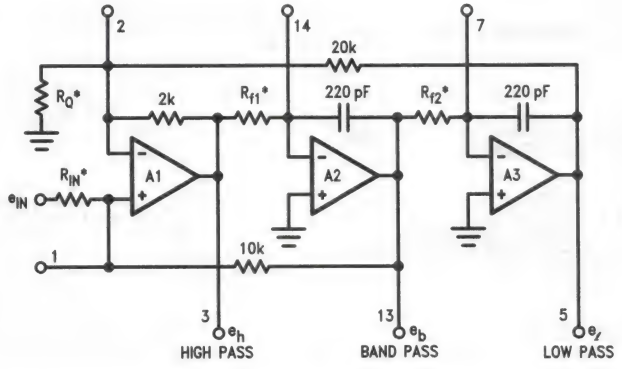
$$\omega_1 = \frac{10^{12}}{R_{f1} \bullet 220}, \quad \omega_2 = \frac{10^{12}}{R_{f2} \bullet 220}$$

$$\omega_0 = \sqrt{0.1\,\omega_1 \omega_2}$$

$$Q = \left[\frac{1 + \frac{10^4}{R_{IN}}}{1.1 + \frac{2 \times 10^3}{R_Q}} \right] \sqrt{0.1 \frac{\omega_2}{\omega_1}} \quad (25)$$

$$R_Q = \frac{2 \times 10^7}{\left(1 + \frac{10^4}{R_{IN}}\right)\left(\frac{\sqrt{0.1 \frac{\omega_2}{\omega_1}}}{Q}\right) - 1.1} \quad (26)$$

Appendix (Continued)



TL/K/10112–32

*External Components

FIGURE 12. Non-Inverting Input ($Q < Q_{MIN}$)

c) Inverting input *(Figure 13)* transfer function equations are:

$$\frac{e_h}{e_{IN}} = \frac{-s^2\left(\frac{2 \times 10^3}{R_{IN}}\right)}{\Delta} \text{ (high pass)} \qquad (27)$$

$$\frac{e_b}{e_{IN}} = \frac{s\,\omega_1\left(\frac{2 \times 10^3}{R_{IN}}\right)}{\Delta} \text{ (band pass)} \qquad (28)$$

$$\frac{e_\ell}{e_{IN}} = \frac{-\omega_1\omega_2\left(\frac{2 \times 10^3}{R_{IN}}\right)}{\Delta} \text{ (low pass)} \qquad (29)$$

$$\omega_1 = \frac{10^{12}}{R_{f1} \bullet 220}, \quad \omega_2 = \frac{10^{12}}{R_{f2} \bullet 220}$$

where

$$\Delta = s^2 + s\,\omega_1\left[\frac{1.1 + \frac{2 \times 10^3}{R_{IN}}}{1 + \frac{10^4}{R_Q}}\right] + 0.1\,\omega_1\,\omega_2 \qquad (30)$$

Appendix (Continued)

$$\left.\frac{e_\ell}{e_{IN}}\right|_{s \to 0} = \frac{2 \times 10^4}{R_{IN}} \text{ (low pass) (DC gain)} \quad (31)$$

$$\left.\frac{e_h}{e_{IN}}\right|_{s \to \infty} = -\frac{2 \times 10^3}{R_{IN}} \begin{matrix}\text{(high pass)} \\ \text{(high freq. gain)}\end{matrix} \quad (32)$$

$$\left.\frac{e_b}{e_{IN}}\right|_{\omega = \omega_0} = \frac{\dfrac{2 \times 10^3}{R_{IN}}\left(1 + \dfrac{10^4}{R_Q}\right)}{1.1 + \dfrac{2 \times 10^3}{R_{IN}}} \begin{matrix}\text{(band pass)} \\ \text{(center freq. gain)}\end{matrix} \quad (33)$$

$$\omega_0 = \sqrt{0.1\ \omega_1\ \omega_2}$$

$$Q = \left[\frac{1 + \dfrac{10^4}{R_Q}}{1.1 + \dfrac{10^4}{R_{IN}}}\right]\sqrt{0.1\,\frac{\omega_2}{\omega_1}} \quad (34)$$

$$R_Q = \frac{10^4}{\dfrac{Q}{\sqrt{0.1\,\dfrac{\omega_2}{\omega_1}}}\left(1.1 + \dfrac{2 \times 10^3}{R_{IN}}\right) - 1} \quad (35)$$

d) Differential input *(Figure 14)* transfer function equations are:

$$\frac{e_h}{e_{IN}} = \frac{s^2\left(\dfrac{2 \times 10^3}{R_{IN2}}\right)}{\Delta} \text{ (high pass)} \quad (36)$$

$$\frac{e_b}{e_{IN}} = \frac{-s\,\omega_1\left(\dfrac{2 \times 10^3}{R_{IN2}}\right)}{\Delta} \text{ (band pass)} \quad (37)$$

$$\frac{e_\ell}{e_{IN}} = \frac{\omega_1\,\omega_2\left(\dfrac{2 \times 10^3}{R_{IN2}}\right)}{\Delta} \text{ (low pass)} \quad (38)$$

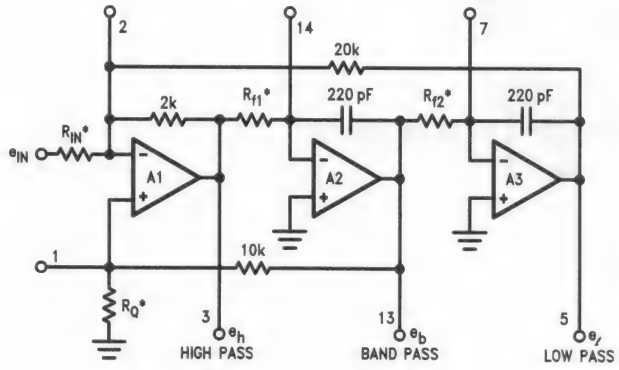
$$\omega_1 = \frac{10^{12}}{R_{f1} \times 220}, \quad \omega_2 = \frac{10^{12}}{R_{f2} \times 220}$$

where

$$\Delta = s^2 + s\,\omega_1\left[\frac{1.1 + \dfrac{2 \times 10^3}{R_{IN2}}}{1 + \dfrac{10^4}{R_Q} + \dfrac{10^4}{R_{IN1}}}\right] + 0.1\,\omega_1\,\omega_2 \quad (39)$$

$$\omega_0 = \sqrt{0.1\ \omega_1\ \omega_2} \quad (40)$$

$$Q = \left[\frac{1 + \dfrac{10^4}{R_Q} + \dfrac{10^4}{R_{IN1}}}{1.1 + \dfrac{2 \times 10^3}{R_{IN2}}}\right]\sqrt{0.1\,\frac{\omega_2}{\omega_1}} \quad (41)$$



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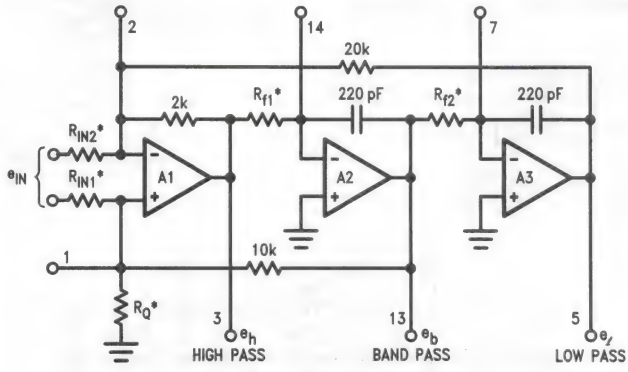
*External Components

FIGURE 13. Inverting Input, Any Q

Appendix (Continued)



*External Components

TL/K/10112–34

FIGURE 14. Differential Input

$$R_Q = \frac{10^4}{\frac{Q}{\sqrt{0.1\,\frac{\omega_2}{\omega_1}}}\left(1.1 + \frac{2\times 10^3}{R_{IN2}}\right) - 1 - \frac{10^4}{R_{IN1}}} \quad (42)$$

e) Notch filter *(Figure 15)* transfer function equations are:

$$\frac{e_n}{e_{IN}} = \frac{(s^2 + \omega_z^2)\left[\frac{1.1}{1 + \frac{R_{IN}}{10^4} + \frac{R_{IN}}{R_Q}}\right]\frac{R_g}{R_h}}{s^2 + s\,\omega_1\left[\frac{1.1}{1 + \frac{10^4}{R_Q} + \frac{10^4}{R_{IN}}}\right] + 0.1\,\omega_1\,\omega_2} \quad (45)$$

$$\omega_1 = \frac{10^{12}}{R_{f1}\times 220'}, \quad \omega_2 = \frac{10^{12}}{R_{f2}\times 220'}, \quad \omega_0 = \sqrt{0.1\,\omega_1\,\omega_2}$$

$$\omega_Z = \omega_0\sqrt{\frac{10\,R_h}{R_L}}$$

$$\left.\frac{e_n}{e_{IN}}\right|_{s\to 0} = \frac{11}{\left(1 + \frac{R_{IN}}{10^4} + \frac{R_{IN}}{R_Q}\right)}\frac{R_g}{R_L}\ \text{(DC gain)} \quad (46)$$

$$\left.\frac{e_n}{e_{IN}}\right|_{s\to\infty} = \frac{1.1}{\left(1 + \frac{R_{IN}}{10^4} + \frac{R_{IN}}{R_Q}\right)}\frac{R_g}{R_h}\ \text{(high freq. gain)} \quad (47)$$

$$\left.\frac{e_n}{e_{IN}}\right|_{\omega=\omega_z} = 0 \quad (48)$$

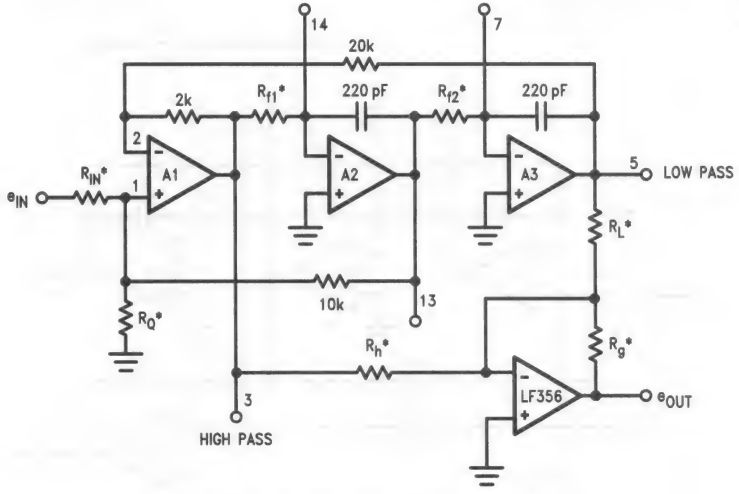


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FIGURE 15. Notch Filter Using an External Amplifier

Appendix (Continued)

f) Input notch filter *(Figure 16)* transfer function equations are:

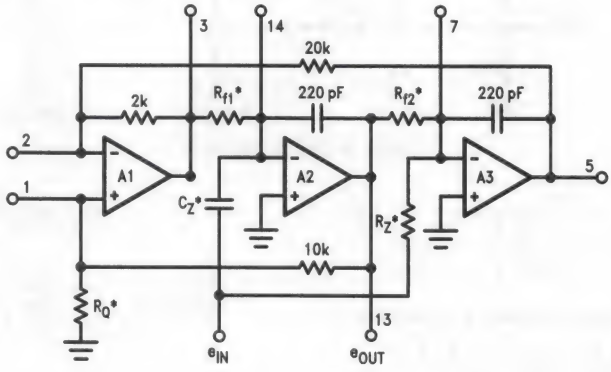
$$\frac{e_{IN}}{e_n} = -\frac{\frac{C_Z}{220 \times 10^{-12}}[s^2 + \omega_Z{}^2]}{s^2 + s\omega_1\left[\frac{1.1\,R_Q}{10^4 + R_Q}\right] + \omega_0{}^2} \quad (49)$$

$$\omega_1 = \frac{10^{12}}{R_{f1} \bullet 220}, \quad \omega_2 = \frac{10^{12}}{R_{f2} \bullet 220}$$

$$\omega_Z = \omega_0\sqrt{\frac{R_{f2} \bullet 220 \times 10^{-12}}{R_Z\,C_Z}}, \quad \omega_0 = \sqrt{0.1\,\omega_1\,\omega_2} \quad (50)$$

$$\left.\frac{e_n}{e_{IN}}\right|_{\omega \to 0} = \frac{-R_{F2}}{R_Z} \quad (51)$$

$$\left.\frac{e_n}{e_{IN}}\right|_{\omega \to \infty} = \frac{C_Z}{220 \times 10^{-12}} \quad (52)$$



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*External Components

FIGURE 16. Input Notch Filter Using 3 Amplifiers

g) All pass *(Figure 17)* transfer function equations are:

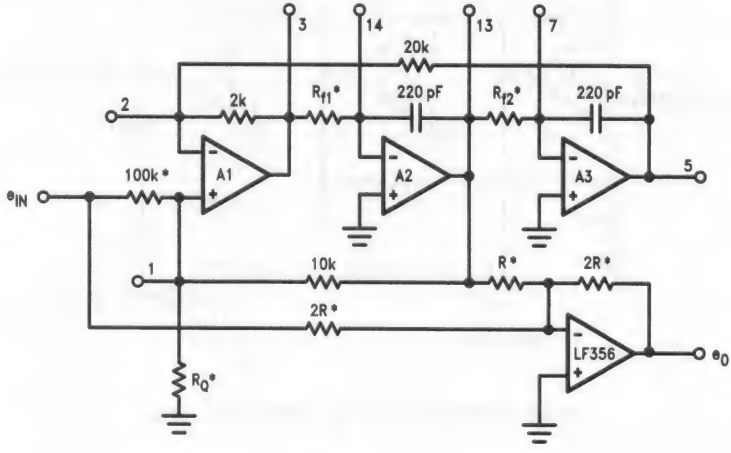
$$\frac{e_o}{e_{IN}} = -\left[\frac{s^2 - s\,\omega_1\left[\frac{1.1}{2 + \frac{R_{IN}}{R_Q}}\right] + \omega_0{}^2}{s^2 + s\,\omega_1\left[\frac{1.1}{2 + \frac{R_{IN}}{R_Q}}\right] + \omega_0{}^2}\right] \quad (53)$$

$$Q = \left[\frac{2 + \frac{10^4}{R_Q}}{1.1}\right]\sqrt{0.1\,\frac{\omega_2}{\omega_1}} \quad (54)$$

$$\omega_1 = \frac{10^{12}}{R_{f1} \bullet 220}, \quad \omega_2 = \frac{10^{12}}{R_{f2} \bullet 220}$$

$$\omega_0 = \sqrt{0.1\,\omega_1\,\omega_2}$$

Time delay at ω_0 is $\frac{2Q}{\omega_0}$ seconds



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*External Components

FIGURE 17. All Pass

Definition of Terms

A_{MAX}	Maximum pass band peak-to-peak ripple
A_{MIN}	Minimum stop band loss
f_z	Frequency of jw axis pole pair
f_o	Frequency of complex pole pair
Q	Quality of pole
f_c	Pass band edge
f_s	Stop band edge
R_f	Pole frequency determining resistance
R_z	Zero Frequency determining resistance
R_Q	Pole quality determining resistance
f_H	Frequency above center frequency at which the gain decreases by 3 dB for a band pass filter
f_L	Frequency below center frequency at which the gain decreases by 3 dB for a band pass filter

Bibliography

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G.S. Moschytz: *"Linear Integrated Networks Design"*, Van Norstrand Reinhold Co., New York, 1975

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National Semiconductor

AF151 Dual Universal Active Filter

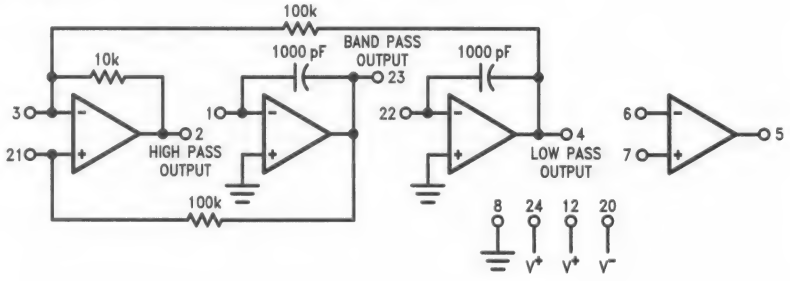
General Description

The AF151 consists of 2 general purpose state variable active filters in a single package. By using only 4 external resistors for each section, various second order functions may be formed. Low pass, high pass and band pass functions are available simultaneously at separate outputs. In addition, there are 2 uncommitted operational amplifiers which are available for buffering or for forming all pass and notch functions. Any of the classical filter configurations, such as Butterworth, Bessel, Cauer and Chebyshev can be easily formed.

Features

- Independent Q, frequency and gain adjustment
- Very low sensitivity to external component variation
- Separate low pass, high pass and band pass outputs
- Operation to 10 kHz
- Q range to 500
- Wide power supply range—±5V to ±18V
- Accuracy—±1%
- Fourth order functions in one package

Circuit Diagrams



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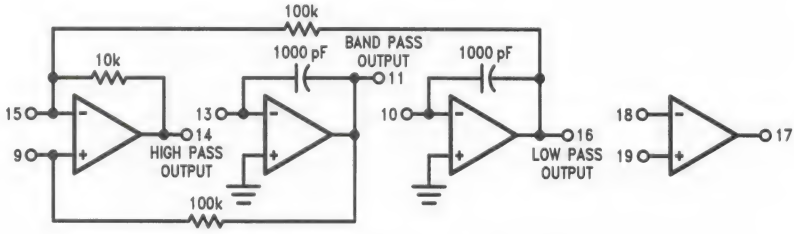

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Order Number AF151-1CJ or AF151-2CJ
See NS Package Number HY24A

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	±18V
Power Dissipation	900 mW/Package
Differential Input Voltage	±36V
Output Short-Circuit Duration (Note 1)	Infinite
Operating Temperature	−25°C to +85°C
Storage Temperature	−25°C to +100°C
Lead Temperature (Soldering, 10 sec.)	300°C

Electrical Characteristics (Complete Active Filter)

Specifications apply for $V_S = \pm 15V$ and over −25°C to +85°C unless otherwise specified.
(Specifications apply for each section.)

Parameter	Conditions	Min	Typ	Max	Units
Frequency Range	$f_c \times Q \leq 50{,}000$			10k	Hz
Q Range	$f_c \times Q \leq 50{,}000$			500	Hz/Hz
f_o Accuracy					
AF151-1C	$f_c \times Q \leq 10{,}000$, $T_A = 25°C$			±2.5	%
AF151-2C	$f_c \times Q \leq 10{,}000$, $T_A = 25°C$			±1.0	
f_o Temperature Coefficient			±50	±150	ppm/°C
Q Accuracy	$f_c \times Q \leq 10{,}000$, $T_A = 25°C$			±7.5	%
Q Temperature Coefficient			±300	±750	ppm/°C
Power Supply Current	$V_S = \pm 15V$		2.5	4.5	mA

Electrical Characteristics (Internal Op Amp) (Note 2)

Parameter	Conditions	Min	Typ	Max	Units
Input Offset Voltage	$R_S \leq 10\ k\Omega$		1.0	6.0	mV
Input Offset Current			4	50	nA
Input Bias Current			30	200	nA
Input Resistance			2.5		MΩ
Large Signal Voltage Gain	$R_L \geq 2k$, $V_{OUT} = \pm 10V$	25	160		V/mV
Output Voltage Swing	$R_L = 10\ k\Omega$	±12	±14		V
	$R_L = 2\ k\Omega$	±10	±13		
Input Voltage Range		±12			V
Common-Mode Rejection Ratio	$R_S \leq 10\ k\Omega$	70	90		dB
Supply Voltage Rejection Ratio	$R_S \leq 10\ k\Omega$	77	96		dB
Output Short-Circuit Current			25		mA
Slew Rate (Unity Gain)			0.6		V/μs
Small Signal Bandwidth			1		MHz
Phase Margin			60		Degrees

Note 1: Any of the amplifiers can be shorted to ground indefinitely; however, more than one should not be simultaneously shorted as the maximum junction temperature will be exceeded.

Note 2: Specifications apply for $V_S = \pm 15V$, $T_A = 25°C$.

Applications Information

The AF151 consists of 2 identical filter sections and 2 uncommitted op amps. The op amps may be used for buffering inputs and outputs, summing amplifiers (for notch filter generation), adjusting gain through the filter sections, additional passive networks to create higher order filters, or simply used elsewhere in the user's system.

The design equations given apply to both sections; however, for clarity, only the pin designations for Section 1 will be shown in the examples and discussion.

See the AF100 datasheet for additional information on this type of filter.

The design equations assume that the user has knowledge of the frequency and Q values for the particular design to be synthesized. If this is not the case, various references and texts are available to help the user in determining these parameters. A bibliography of recommended texts can also be found in the AF100 datasheet.

CIRCUIT DESCRIPTION AND OPERATION

A schematic of one section of the AF151 is shown in *Figure 1*. Amplifier A1 is a summing amplifier with inputs from integrator A2 to the non-inverting input and integrator A3 to the inverting input. Amplifier A4 is an uncommitted amplifier.

By adding external resistors the circuit can be used to generate the second order system.

$$T(s) = \frac{a_3 s^2 + a_2 s + a_1}{s^2 + b_2 s + b_1}$$

The denominator coefficients determine the complex pole pair location and the quality of the poles where

$\omega_o = \sqrt{b_1}$ = the radian center frequency

$Q = \frac{\omega_o}{b_2}$ = the quality of the complex pole pair

If the output is taken from the output of A1, numerator coefficients a_1 and a_2 equal zero, and the transfer function becomes:

$$T(s) = \frac{a_3 s^2}{s^2 + \frac{\omega_o}{Q} + \omega_o^2} \quad \text{(High Pass)}$$

If the output is taken from the output of A2, numerator coefficients a_1 and a_3 equal zero and the transfer function becomes:

$$T(s) = \frac{a_2 s}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} \quad \text{(Band Pass)}$$

If the output is taken from the output of A3, numerator coefficients a_3 and a_2 equal zero and the transfer function becomes:

$$T(s) = \frac{a_1}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} \quad \text{(Low Pass)}$$

Using proper input and output connections the circuit can also be used to generate the transfer functions for a notch and all pass filter.

In the transfer function for a notch function a_2 becomes zero, a_1 equals ω_z^2 and a_3 equals 1. The transfer function becomes:

$$T(s) = \frac{s^2 + \omega_z^2}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} \quad \text{(Notch)}$$

In the all pass transfer function $a_1 = \omega_o^2$, $a_2 = -\omega_o/Q$ and $a_3 = 1$. The transfer function becomes:

$$T(s) = \frac{s^2 - \frac{\omega_o}{Q} s + \omega_o^2}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} \quad \text{(All Pass)}$$

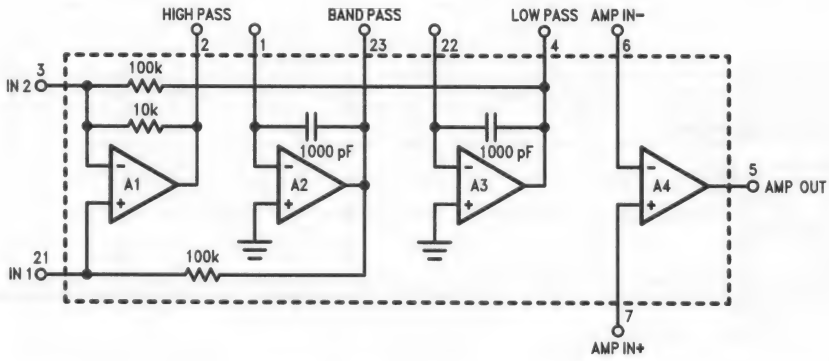

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FIGURE 1. AF151 Schematic (Section 1)

Applications Information (Continued)

FREQUENCY CALCULATIONS

For operation above 200 Hz, the frequency of each section of the AF151 is set by 2 equal valued resistors. These resistors couple the output of the first op amp (pin 2) to the input of the second op amp (pin 1) and the output of the second op amp (pin 23) to the input of the third op amp (pin 22).

The value for R_f is given by:

$$R_f = \frac{50.33 \times 10^6}{f_o}\Omega \quad (1)$$

For operation below 200 Hz, "T" tuning should be used as shown in *Figure 3*.

For this configuration,

$$R_S = \frac{R_T{}^2}{R_f - 2R_T} \quad (2)$$

where R_T or R_S can be chosen arbitrarily, once R_f is found from Equation 1.

Q CALCULATIONS

To set the Q of each section of the AF151, one resistor is required. The value of the Q setting resistor depends on the input connection (inverting or non-inverting) and the input resistance. Because the input resistance does affect the Q, it is often desirable to use one of the uncommitted op amps to provide a buffer between the signal source impedance and the input resistor used to set the Q.

To determine which connection is required for a particular Q, arbitrarily select a value of R_{IN} *(Figure 4)* and calculate Q_{MIN} according to Equation 3.

$$Q_{MIN} = \frac{1 + \frac{10^5}{R_{IN}}}{3.48} \quad (3)$$

If the Q required for the circuit is greater than Q_{MIN}, use Equation 4 to calculate the value of R_Q and the connection shown in *Figure 4*.

$$R_Q = \frac{10^5}{3.48Q - 1 - \frac{10^5}{R_{IN}}} \quad (4)$$

If the Q required for the circuit is less than Q_{MIN}, use Equation 5 to calculate the value of R_Q and the connection shown in *Figure 5*.

$$R_Q = \frac{10^4}{\frac{0.3162}{Q}\left(1 + \frac{10^5}{R_{IN}}\right) - 1.1} \quad (5)$$

Both connections shown in *Figures 4* and *5* are "non-inverting" relative to the phase relationship between the input signal and the low pass output.

For any Q, Equation 6 may be used with the "inverting" connection shown in *Figure 6*.

$$R_Q = \frac{10^5}{3.16Q\left(1.1 + \frac{10^4}{R_{IN}}\right) - 1} \quad (6)$$

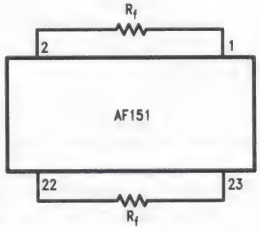

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FIGURE 2. Frequency Tuning

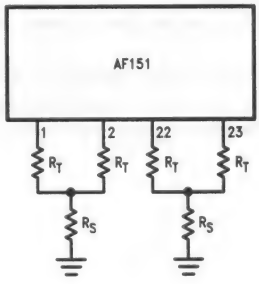

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FIGURE 3. "T" Tuning for Low Frequency

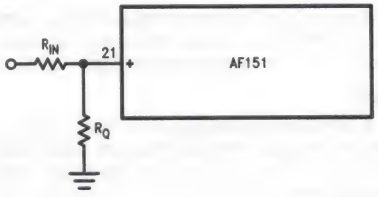

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FIGURE 4. Connection for $Q > Q_{MIN}$

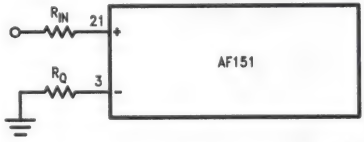

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FIGURE 5. Connection for $Q < Q_{MIN}$

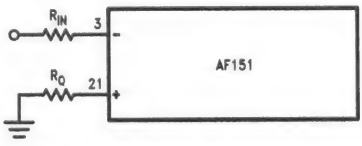

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FIGURE 6. Connection for Any Q, Inverting

Applications Information (Continued)

NOTCH TUNING

When the low pass output and the high pass output are summed together, the result is a notch *(Figure 7)*.

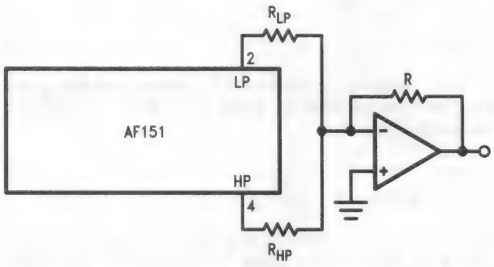

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FIGURE 7. Notch Filter

The relationship between R_{LP}, R_{HP}, f_o and f_z, the location of the notch, is given by Equation 7.

$$R_{HP}\left(\frac{f_z}{f_o}\right)^2\frac{R_{LP}}{10} \quad (7)$$

Again, it is advantageous to use one of the uncommitted op amps to perform this summing function to prevent loading of this stage or the resistors R_{LP} and R_{HP} from effecting the Q of subsequent stages. Resistor R can be used to set the gain of the filter section.

GAIN CALCULATIONS

The following list of equations will be helpful in calculating the relationship between the external components and various important parameters. The following definitions are used:

A_L — Gain from input to low pass output at DC

A_H — Gain from input to high pass output at high frequency

A_B — Gain from input to band pass output at center frequency

For *Figure 4:*

$$A_L = \frac{11}{\Delta}$$

$$A_H = \frac{1.1}{\Delta}$$

$$A_B = \frac{-\left(1+\frac{10^5}{R_Q}+\frac{10^5}{R_{IN}}\right)}{\Delta}$$

$$\Delta = 1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{R_Q}$$

For *Figure 5:*

$$A_L = \frac{11+\frac{10^5}{R_Q}}{\Delta}$$

$$A_H = \frac{1.1+\frac{10^4}{R_Q}}{\Delta}$$

$$AB = \frac{-\left(1+\frac{10^5}{R_{IN}}\right)}{\Delta}$$

$$\Delta = 1+\frac{R_{IN}}{10^5}$$

For *Figure 6:*

$$A_L = -\frac{10^5}{R_{IN}}$$

$$A_H = -\frac{10^4}{R_{IN}}$$

$$A_B = \frac{\frac{10^5}{R_{IN}}\left(1+\frac{10^5}{R_Q}\right)}{11+\frac{10^5}{R_{IN}}}$$

For *Figure 7:*

At low frequency, when $f_o < f_z$, the gain to the output of the summing op amp is:

$$A_L = \frac{11\left(\frac{R}{R_{LP}}\right)}{\left(1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{R_Q}\right)}$$

At high frequency, when $f_o > f_z$, the gain to the output of the summing op amp is:

$$A_H = \frac{1.1\left(\frac{R}{R_{HP}}\right)}{\left(1+\frac{R_{IN}}{10^5}+\frac{R_{IN}}{R_Q}\right)}$$

At the notch, ideally the gain is zero (0).

TUNING TIPS

In applications where 2% to 3% accuracy is not sufficient to provide the required filter response, the AF151 stages can be tuned by adding trim pots or trim resistors in series or parallel with one of the frequency determining resistors and the Q determining resistor.

When tuning a filter section, no matter what output configuration is to be used in the circuit, measurements are made between the input and the band pass output.

Before any tuning is attempted, the low pass output should be checked to see that the output is not clipping. At the center frequency of the section, the low pass output is 10 dB higher than the band pass output and 20 dB higher than the high pass. This should be kept in mind because if clipping occurs, the results obtained when tuning will be incorrect.

Frequency Tuning

By adjusting resistor R_f, center frequency of a section can be adjusted. Adjusting center frequency by phase is the most accurate but tuning for maximum gain is also correct.

Q Tuning

The Q is tuned by adjusting the R_Q resistor. To tune the Q correctly, the signal source must have an output impedance very much lower than the input resistance of the filter since the input resistance affects the Q. The input must be driven through the same resistance the circuit will "see" to obtain precise adjustment.

Applications Information (Continued)

The lower 3 dB (45°) frequency, f_L, and the upper 3 dB (45°) frequency, f_H, can be calculated by the following equations:

$$f_H = \left(\frac{1}{20} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right) \times \left(f_o\right)$$

where f_o = center frequency

$$f_L = \left(\sqrt{\left(\frac{1}{2Q}\right)^2 + 1} - \frac{1}{2Q}\right) \times \left(f_o\right)$$

When adjusting the Q, set the signal source to either f_H or f_L and adjust for 45°C phase change or a 3 dB gain change.

Notch Tuning

If a circuit has a jw axis zero pair, the notch can be tuned by adjusting the ratio of the summing resistors (low pass/high pass summing).

In either case, the signal is connected to the input and the proper resistor is adjusted for a null at the output.

TUNING PROCEDURE

Center Frequency Tuning

Set oscillator to center frequency desired for the filter section, adjust amplitude and check that clipping does not occur at the low pass output.

Adjust the R_f resistor until the phase shift between input and band pass output is 180° or 0°, depending upon the connection.

Q Tuning

Set oscillator to upper or lower 45° frequency (see tuning tips) and tune the Q resistor until the phase shift is 135° (upper 45° frequency) or 225° (lower 45° frequency).

Zero Tuning (Notch Tuning)

Set the oscillator output to the zero frequency and tune one of the summing resistors for a null at the output of the summing amplifier.

Gain Adjust

Set the oscillator to any desired frequency and the gain can be adjusted by measuring the output of the summing amplifier and adjusting the feedback resistance.

DESIGN EXAMPLE

Assume 2 band pass filters are required to separate FSK data.

f1 = 800 Hz, Q = 40

f2 = 1000 Hz, Q = 50

The gain through each filter is to be 10 V/V (20 dB).

Since the design is similar for both sections, only the first section design will be shown for the example.

(a) From Equation 1

$$R_f = \frac{50.33 \times 10^6}{f_o} = \frac{50.33 \times 10^6}{800}$$

$R_f = 62.9k$

(b) Checking Q_{MIN} from Equation 3, arbitrarily let R_{IN} = 300k.

$$Q_{MIN} = \frac{1 + \frac{10^5}{R_{IN}}}{3.48} = \frac{1 + \frac{10^5}{3 \times 10^5}}{3.48} = 0.383$$

Since the Q required for the design (Q = 40), is greater than Q_{MIN}, the circuit of *Figure 4* or *Figure 6* may be used. Arbitrarily we shall select the circuit of *Figure 4*.

(c) From Equation 4, R_Q is found to be

$$R_Q = \frac{10^5}{3.48Q - 1 - \frac{10^5}{R_{IN}}} = \frac{10^5}{(3.48)(40) - 1 - \frac{10^5}{3 \times 10^5}}$$

or $R_Q = 725\Omega$

(d) Calculate the center frequency gain for *Figure 4*.

$$A_B = \frac{-\left(1 + \frac{10^5}{R_Q} + \frac{10^5}{R_{IN}}\right)}{\left(1 + \frac{R_{IN}}{10^5} + \frac{R_{IN}}{R_Q}\right)} = \frac{-(1 + 137.9 + 0.333)}{(1 + 3.0 + 414)}$$

A_B = 0.333 V/V

Since the gain at f_o is 0.333 V/V, a gain of 10 V/V can be obtained by using the uncommitted operational amplifier with a gain of 30.03 as shown in *Figure 8*.

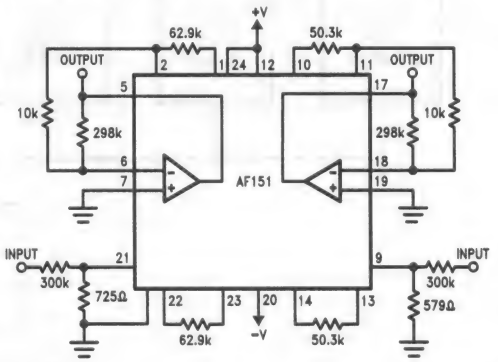

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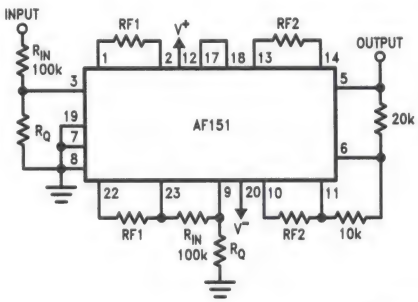
FIGURE 8. Dual Band Pass Filter



TL/K/10113–11

FIGURE 9. Telephone Multifrequency (MF) Band Pass Filter

Applications Information (Continued)

FREQ	BW	f_c	f1	Q1 & Q2	f2	RF1	RF2	RQ
700	75	698.4	665.6	17	732.8	75.62k	68.68k	1.749k
900	75	898.7	865.8	21.8	932.9	58.13k	53.95k	1.354k
1100	75	1098.8	1065.7	26.7	1132.9	47.23k	44.43k	1.100k
1300	75	1298.9	1265.8	31.6	1332.9	39.76k	37.76k	926.2Ω
1500	75	1499.0	1465.8	36.4	1532.9	34.34k	32.83k	802.1Ω
1700	75	1699.1	1665.9	41.3	1733.0	30.21k	29.04k	705.6Ω

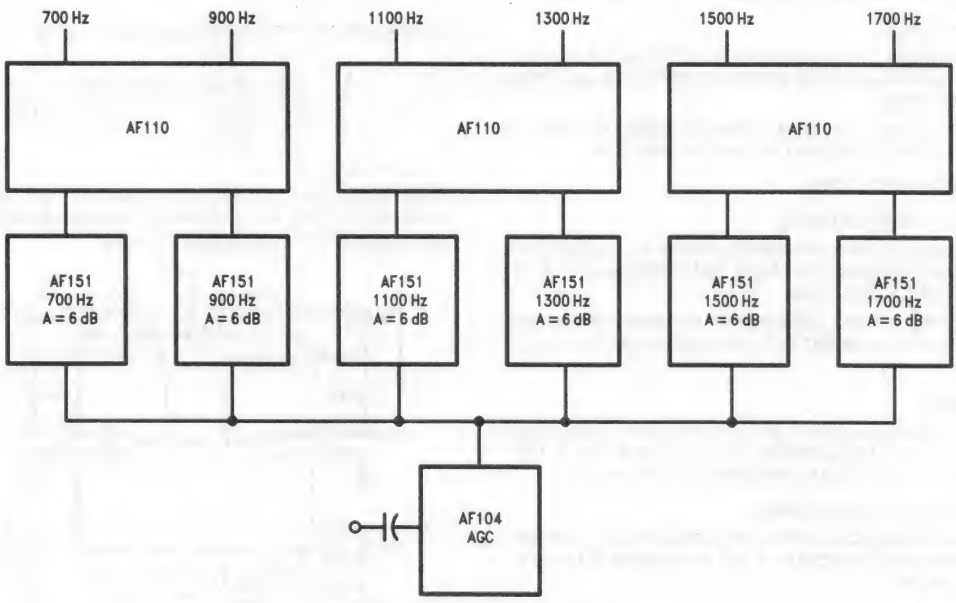


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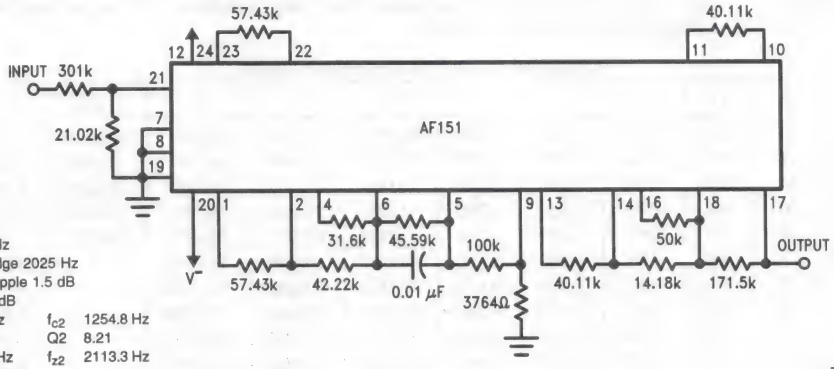
FIGURE 10. MF Tone Receiver

Applications Information (Continued)



Cutoff 1270 Hz
Stop Band Edge 2025 Hz
Band Pass Ripple 1.5 dB
Rejection 59 dB

f_{c1}	876.3 Hz	f_{c2}	1254.8 Hz
Q1	1.75	Q2	8.21
f_{z1}	3201.7 Hz	f_{z2}	2113.3 Hz
f_R	356.9 Hz		

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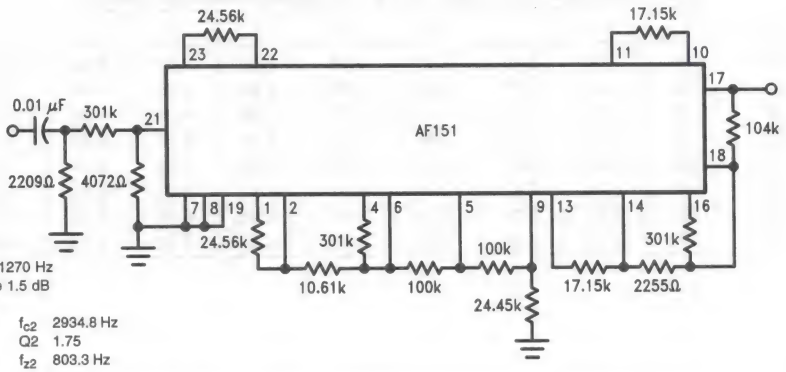
FIGURE 11. Low Pass Low Speed Asynchronous FSK Modem Filter



Cutoff 2025 Hz
Stop Band Edge 1270 Hz
Band Pass Ripple 1.5 dB
Rejection 59 dB

f_{c1}	2049.6 Hz	f_{c2}	2934.8 Hz
Q1	8.21	Q2	1.75
f_{z1}	1216.9 Hz	f_{z2}	803.3 Hz
f_R	7206 Hz		

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FIGURE 12. High Pass Low Speed Asynchronous FSK Modem Filter

Applications Information (Continued)

Standard Resistance Values are obtained from the Decade Table by multiplying by multiples of 10. As an example, 1.33 can represent 1.33Ω, 133Ω, 1.33 kΩ, 13.3 kΩ, 133 kΩ, 1.33 MΩ.

Standard 5% and 2% Resistance Values

Ω	Ω	Ω	Ω	Ω	Ω	Ω	Ω	Ω	Ω	Ω	MΩ	
10	27	68	180	470	1,200	3,300	8,200	22,000	56,000	150,000	0.24	0.62
11	30	75	200	510	1,300	3,600	9,100	24,000	62,000	160,000	0.27	0.68
12	33	82	220	560	1,500	3,900	10,000	27,000	68,000	180,000	0.30	0.75
13	36	91	240	620	1,600	4,300	11,000	30,000	75,000	200,000	0.33	0.82
15	39	100	270	680	1,800	4,700	12,000	33,000	82,000	220,000	0.36	0.91
16	43	110	300	750	2,000	5,100	13,000	36,000	91,000		0.39	1.0
18	47	120	330	820	2,200	5,600	15,000	39,000	100,000		0.43	1.1
20	51	130	360	910	2,400	6,200	16,000	43,000	110,000		0.47	1.2
22	56	150	390	1,000	2,700	6,800	18,000	47,000	120,000		0.51	1.3
24	62	160	430	1,100	3,000	7,500	20,000	51,000	130,000		0.56	1.5

Decade Table Determining ½% and 1% Standard Resistance Values

Ω	Ω	Ω	Ω	Ω	Ω	Ω	Ω	Ω	Ω	Ω	MΩ
1.00	1.21	1.47	1.78	2.15	2.61	3.16	3.83	4.64	5.62	6.81	8.25
1.02	1.24	1.50	1.82	2.21	2.67	3.24	3.92	4.75	5.76	6.98	8.45
1.05	1.27	1.54	1.87	2.26	2.74	3.32	4.02	4.87	5.90	7.15	8.66
1.07	1.30	1.58	1.91	2.32	2.80	3.40	4.12	4.99	6.04	7.32	8.87
1.10	1.33	1.62	1.96	2.37	2.87	3.48	4.22	5.11	6.19	7.50	9.09
1.13	1.37	1.65	2.00	2.43	2.94	3.57	4.32	5.23	6.34	7.68	9.31
1.15	1.40	1.69	2.05	2.49	3.01	3.65	4.42	5.36	6.49	7.87	9.53
1.18	1.43	1.74	2.10	2.55	3.09	3.74	4.53	5.49	6.65	8.06	9.76

National Semiconductor

LMF90
4th-Order Elliptic Notch Filter

General Description

The LMF90 is a fourth-order elliptic notch (band-reject) filter based on switched-capacitor techniques. No external components are needed to define the response function. The depth of the notch is set using a two-level logic input, and the width is programmed using a three-level logic input. Two different notch depths and three different ratios of notch width to center frequency may be programmed by connecting these pins to V^+, ground, or V^-. Another three-level logic pin sets the ratio of clock frequency to notch frequency.

An internal crystal oscillator is provided. Used in conjunction with a low-cost color TV crystal and the internal clock frequency divider, a notch filter can be built with center frequency at 50 Hz, 60 Hz, 100 Hz, 120 Hz, 150 Hz, or 180 Hz for rejection of power line interference. Several LMF90s can be operated from a single crystal. An additional input is provided for an externally-generated clock signal.

Features

- Center frequency set by external clock or on-board clock oscillator
- No external components needed to set response characteristics
- Notch width, attenuation, and clock-to-center-frequency ratio independently programmable
- 14 pin 0.3″ wide package

Key Specifications

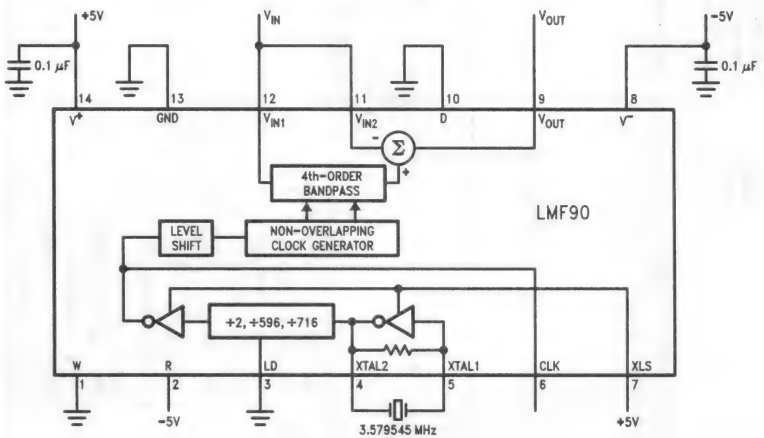
- f_0 Range — 0.1 Hz to 30 kHz
- f_0 accuracy over full temperature range (max) — 1.5%
- Supply voltage range — ±2V to ±7.5V or 4V to 15V
- Passband Ripple (typ) — 0.25 dB
- Attenuation at f_0 (typ) — 39 dB or 48 dB (selectable)
- f_{CLK}: f_0 — 100:1, 50:1, or 33.3:1
- Notch Bandwidth (typ) — 0.127 f_0, 0.26 f_0, or 0.55 f_0
- Output offset voltage (max) — 120 mV

Applications

- Automatic test equipment
- Communications
- Power line interference rejection

Typical Connection

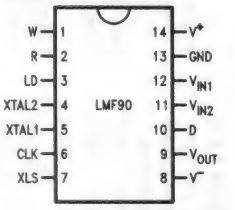
60 Hz Notch Filter



TL/H/10354–1

Connection Diagram

Dual-In-Line Package



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Top View

Order Number LMF90CCN, LMF90CCWM, LMF90CIJ or LMF90CMJ
See NS Package Number J14A, M14B or N14A

Absolute Maximum Ratings (Notes 1 & 3)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage ($V_S = V^+ - V^-$)	−0.3V to +16V
Voltage at any Input or Output	V^- −0.3V to V^+ +0.3V
Input Current at any Pin (Note 10)	5 mA
Package Input Current (Note 10)	20 mA
Power Dissipation (Note 5)	500 mW
ESD Susceptability (Note 6)	
Pin 9	1800V
All Other Pins	2000V
Soldering Information (Note 4)	
N Package (Soldering, 10 sec.)	260°C
J Package (Soldering, 10 sec.)	300°C
Storage Temperature Range	−65°C to +150°C
Junction Temperature	150°C

Operating Ratings (Notes 2 & 3)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
LMF90CCN, LMF90CCWM	$0°C \le T_A \le +70°C$
LMF90CIJ	$-40°C \le T_A \le +85°C$
LMF90CMJ	$-55°C \le T_A \le +125°C$
Supply Voltage Range	4.0V to 15.0V

AC Electrical Characteristics

The following specifications apply for $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for $T_A = T_{MIN}$ to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Symbol	Parameter	Conditions	LMF90CCN, LMF90CCWM			LMF90CIJ, LMF90CMJ			Units (Limit)
			Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	
f_O	Center Frequency Range		0.1	30	**30**	0.1	**30**		Hz (Min) kHz (Max)
f_{CLK}	Clock Frequency Range	Pin 6 Pin 6 Pins 4 and 5	10	1.5 4.0	**1.5** **4.0**	10	**1.5** **4.0**		Hz (Min) MHz (Max) MHz (Max)
f_{CLK}/f_{O1}	Clock-to-Center-Frequency Ratio	$W = D = V^-$, $R = V^+$, f_{CLK} = 167 kHz		33.5 ±1%	**33.5 ±1.5%**		**33.5 ±1.5%**		(Max)
f_{CLK}/f_{O2}		$W = D = R$ = GND, f_{CLK} = 250 kHz		50.25 ±1%	**50.25 ±1.5%**		**50.25 ±1.5%**		(Max)
f_{CLK}/f_{O3}		$W = V^+$, D = GND, $R = V^-$, f_{CLK} = 500 kHz		100.5 ±1%	**100.5 ±1.5%**		**100.5 ±1.5%**		(Max)
H_{ON}	Passband Gain	DC and 20 kHz, $W = D = V^-$, $R = V^+$, f_{CLK} = 167 kHz	0	±0.2	**±0.2**	0	**±0.2**		dB (Max)
		$W = D = R$ = GND, f_{CLK} = 250 kHz	0	±0.2	**±0.2**	0	**±0.2**		dB (Max)
		$W = V^+$, D = GND, $R = V^-$, f_{CLK} = 500 kHz	0	±0.2	**±0.2**	0	**±0.2**		dB (Max)

AC Electrical Characteristics

The following specifications apply for $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for $T_A = T_{MIN}$ to T_{MAX};** all other limits $T_A = T_J = 25°C$. (Continued)

Symbol	Parameter	Conditions	LMF90CCN, LMF90CCWM			LMF90CIJ, LMF90CMJ			Units (Limit)
			Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	
PBW	Ratio of Passband Width to Center Frequency	W = D = V^-, R = V^+, f_{CLK} = 167 kHz		0.1275 ±0.0175	**0.1275 ±0.0175**		**0.1275 ±0.0175**		(Max)
		W = D = R = GND, f_{CLK} = 250 kHz		0.265 ±0.025	**0.265 ±0.025**		**0.265 ±0.025**		(Max)
		W = V^+, D = GND, R = V^-, f_{CLK} = 500 kHz		0.550 ±0.05	**0.550 ±0.05**		**0.550 ±0.05**		(Max)
$A_{Min1}@f_{O1}$	Gain at Center Frequency	W = D = V^-, R = V^+, f_{CLK} = 167 kHz	−39	−30	**−30**	−39	**−30**		dB (Max)
$A_{Min2}@f_{O2}$		W = D = R = GND, f_{CLK} = 250 kHz	−48	−36.5	**−36.5**	−48	**−36.5**		dB (Max)
$A_{Min3}@f_{O3}$		W = V^+, D = GND, R = V^-, f_{CLK} = 500 kHz	−48	−36.5	**−36.5**	−48	**−36.5**		dB (Max)
	Additional Center Frequency Gain Tests at f_{O1}	W = GND, D = V^-, R = V^+, f_{CLK} = 167 kHz	−36	−30	**−30**	−36	**−30**		dB (Max)
		W = V^+, D = V^-, R = V^+, f_{CLK} = 167 kHz	−36	−30	**−30**	−36	**−30**		dB (Max)
		W = V^-, D = GND, R = V^+, f_{CLK} = 167 kHz	−42	−30	**−30**	−42	**−30**		dB (Max)
		W = D = GND, R = V^+, f_{CLK} = 167 kHz	−48	−35	**−35**	−48	**−35**		dB (Max)
		W = V^+, D = GND,R = V^+, f_{CLK} = 167 kHz	−48	−35	**−35**	−48	**−35**		dB (Max)

AC Electrical Characteristics

The following specifications apply for $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for $T_A = T_{MIN}$ to T_{MAX};** all other limits $T_A = T_J = 25°C$. (Continued)

Symbol	Parameter	Conditions		LMF90CCN, LMF90CCWM			LMF90CIJ, LMF90CMJ			Units (Limit)
				Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	
	Additional Center Frequency Gain Tests at f_{O2}	$W = V^-, D = V^-, R = GND$, $f_{CLK} = 250$ kHz		−36	−30	**−30**	−36	**−30**		dB (Max)
		$W = GND, D = V^-, R = GND$, $f_{CLK} = 250$ kHz		−36	−30	**−30**	−36	**−30**		dB (Max)
		$W = V^+, D = V^-, R = GND$, $f_{CLK} = 250$ kHz		−36	−30	**−30**	−36	**−30**		dB (Max)
		$W = V^-, D = R = GND$, $f_{CLK} = 250$ kHz		−42	−30	**−30**	−42	**−30**		dB (Max)
		$W = V^+, D = R = GND$, $f_{CLK} = 250$ kHz		−48	−35	**−35**	−48	**−35**		dB (Max)
	Additional Center Frequency Gain Tests at f_{O3}	$W = D = R = V^-$, $f_{CLK} = 500$ kHz		−36	−30	**−30**	−36	**−30**		dB (Max)
		$W = GND, D = V^-, R = V^-$, $f_{CLK} = 500$ kHz		−36	−30	**−30**	−36	**−30**		dB (Max)
		$W = V^+, D = V^-, R = V^-$, $f_{CLK} = 500$ kHz		−36	−30	**−30**	−36	**−30**		dB (Max)
		$W = V^-, D = GND, R = V^-$, $f_{CLK} = 500$ kHz		−42	−30	**−30**	−42	**−30**		dB (Max)
		$W = D = GND, R = V^-$, $f_{CLK} = 500$ kHz		−48	−35	**−35**	−48	**−35**		dB (Max)
A_{3a}	Gain at $f_3 = 0.995\ f_{O1}$	$W = D = V^-, R = V^+$, $f_{CLK} = 167$ kHz		−41	−30	**−30**	−41	**−30**		dB (Max)
A_{4a}	Gain at $f_4 = 1.005\ f_{O1}$			−41	−30	**−30**	−41	**−30**		dB (Max)
A_{3b}	Gain at $f_3 = 0.992\ f_{O2}$	$W = D = R = GND$, $f_{CLK} = 250$ kHz		−40	−35	**−35**	−40	**−35**		dB (Max)
A_{4b}	Gain at $f_4 = 1.008\ f_{O2}$			−40	−35	**−35**	−40	**−35**		dB (Max)
A_{3c}	Gain at $f_3 = 0.982\ f_{O3}$	$W = V^+, D = GND, R = V^-$		−41	−35	**−35**	−41	**−35**		dB (Max)
A_{4c}	Gain at $f_4 = 1.018\ f_{O3}$	$f_{CLK} = 500$ kHz		−41	−35	**−35**	−41	**−35**		dB (Max)
A_{max1}	Passband Ripple	$W = D = V^-, R = V^+$, $f_{CLK} = 167$ kHz	$f_5 = 0.914\ f_{O1}$	0.25	0.9	**0.9**	0.25	**0.9**		dB (Max)
				0.25	0	**0**	0.25	**0**		dB (Min)
			$f_6 = 1.094\ f_{O1}$	0.25	0.9	**0.9**	0.25	**0.9**		dB (Max)
				0.25	0	**0**	0.25	**0**		dB (Min)

AC Electrical Characteristics

The following specifications apply for $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for $T_A = T_{MIN}$ to T_{MAX};** all other limits $T_A = T_J = 25°C$. (Continued)

Symbol	Parameter	Conditions		LMF90CCN, LMF90CCWM			LMF90CIJ, LMF90CMJ			Units (Limit)
				Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	
A_{Max2}	Passband Ripple	W = D = R = GND, f_{CLK} = 250 kHz	$f_5 = 0.830\ f_{O2}$	0.25 0.25	0.9 0	**0.9** **0**	0.26 0.25	**0.9** **0**		dB (Max) dB (Min)
			$f_6 = 1.205\ f_{O2}$	0.25 0.25	0.9 0	**0.9** **0**	0.25 0.25	**0.9** **0**		dB (Max) dB (Min)
A_{Max3}	Passband Ripple	W = V^+, D = GND, R = V^- f_{CLK} = 500 kHz	$f_5 = 0.700\ f_{O3}$	0.25 0.25	0.9 0	**0.9** **0**	0.25 0.25	**0.9** **0**		dB (Max) dB (Min)
			$f_6 = 1.428\ f_{O3}$	0.25 0.25	0.9 0	**0.9** **0**	0.25 0.25	**0.9** **0**		dB (Max) dB (Min)
E_n	Output Noise	20 kHz Bandwidth W = D = V^-, R = V^+, f_{CLK} = 167 kHz W = D = R = GND, f_{CLK} = 250 kHz W = V^+, D = GND, R = V^-, f_{CLK} = 500 kHz		670 370 250			670 370 250			μVrms μVrms μVrms
	Clock Feedthrough			50			50			mVp−p
GBW	Output Buffer Gain Bandwidth			1			1			MHz
SR	Output Buffer Slew Rate			3			3			V/μs
C_L	Maximum Capacitive Load			200			200			pF

DC Electrical Characteristics

The following specifications apply for $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface Limits Apply for $T_A = T_{MIN}$ to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Symbol	Parameter	Conditions	LMF90CCN, LMF90CCWM			LMF90CIJ, LMF90CMJ			Units (Limit)
			Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Typ (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	
I_S	Power Supply Current	f_{CLK} = 500 kHz, $V_{IN1} = V_{IN2}$ = GND	2.35	5.0	**5.0**	2.35	**5.0**		mA (Max)
V_{OS}	Output Offset Voltage	W = D = V^-, R = V^+, f_{CLK} = 167 kHz W = D = R = GND, f_{CLK} = 250 kHz W = V^+, D = GND, R = V^-, f_{CLK} = 500 kHz	±50 ±60 ±80	±120 ±140 ±170	**±120** **±140** **±170**	±50 ±60 ±80	**±120** **±140** **±170**		mV (Max) mV (Max) mV (Max)
V_{OUT}	Output Voltage Swing	R_L = 5 kΩ	+4.2, −4.7	±4.0	**±4.0**	+4.2, −4.7	**±4.0**		V (Min)
V_{I1}	Logical "Low" Input Voltage	Pins 1, 2, 3, 7, and 10		−4.0	**−4.0**		**−4.0**		V (Max)
V_{I2}	Logical "GND" Input Voltage	Pins 1, 2, 3, 7, and 10		+1.0 −1.0	**+1.0** **−1.0**		**+1.0** **−1.0**		V (Max) V (Min)
V_{I3}	Logical "High" Input Voltage	Pins 1, 2, 3, and 7		+4.0	**+4.0**		**+4.0**		V (Min)
I_{IN}	Input Current	Pins 1, 2, 3, 7, and 10		±10	**±10**		**±10**		μA (Max)
V_{IL}	Logical "0" Input Voltage, Pins 5 and 6	Pin 5, XLS = V^+ or Pin 6, XLS = GND		−4.0	**−4.0**		**−4.0**		V (Max)
V_{IH}	Logical "1" Input Voltage, Pins 5 and 6			+4.0	**+4.0**		**+4.0**		V(Min)
V_{IL}	Logical "0" Input Voltage, Pin 6	$V^+ - V^- = 10V$, XLS = V^- or $V^+ = +5V$, $V^- = 0V$, XLS = +2.5V		+0.8	**+0.8**		**+0.8**		V (Max)
V_{IH}	Logical "1" Input Voltage, Pin 6			+2.0	**+2.0**		**+2.0**		V (Min)
V_{OL}	Logical "0" Output Voltage, Pin 6	XLS = V^+, $\|I_{OUT}\|$ = 4 mA		−4.0	**−4.0**		**−4.0**		V (Max)
V_{OH}	Logical "1" Output Voltage, Pin 6			+4.0	**+4.0**		**+4.0**		V (Min)

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur.

Note 2: Operating Ratings indicate conditions for which the device is intended to be functional. These ratings do not guarantee specific performance limits, however. For guaranteed specifications and test conditions, see the Electrical Characteristics. The guaranteed specifications apply only for the test conditions listed. Some performance characteristics may degrade when the device is not operated under the listed test conditions.

Note 3: All voltages are measured with respect to GND unless otherwise specified.

Note 4: See AN450 "Surface Mounting Methods and Their Effect on Product Reliability" or the section titled "Surface Mount" found in any current Linear Data Book for other methods of soldering surface mount devices.

Note 5: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, Θ_{JA} and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\Theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 150°C$, and the typical thermal resistance (Θ_{JA}) when board mounted is 61°C/W for the LMF90CCN, 134°C/W for the LMF90CCWM, and 59°C/W for the LMF90CIJ and CMJ.

Note 6: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 7: Typicals are at $T_J = 25°C$ and represent the most likely parametric norm.

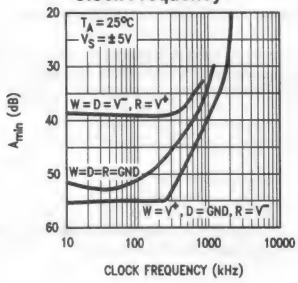
Note 8: Tested Limits are guaranteed and 100% tested.

Note 9: Design Limits are guaranteed, but not 100% tested.

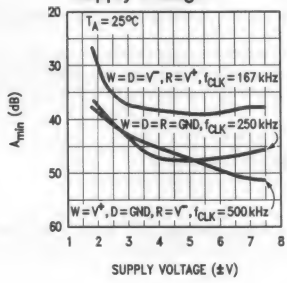
Note 10: When the input voltage (V_{IN}) at any pin exceeds the power supplies ($V_{IN} < V^-$ or $V_{IN} > V^+$), the current at that pin should be limited to 5 mA. The 20 mA maximum package input current rating limits the number of pins that can safely exceed the power supplies with an input current of 5 mA to four.

Typical Performance Characteristics

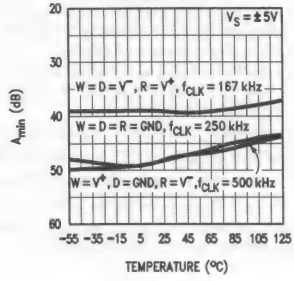
Notch Depth vs Clock Frequency



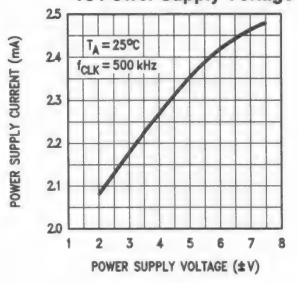
Notch Depth vs Supply Voltage



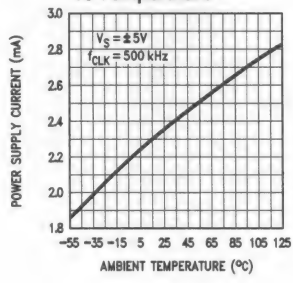
Notch Depth vs Temperature



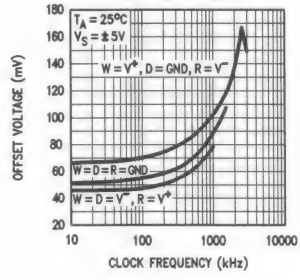
Power Supply Current vs Power Supply Voltage



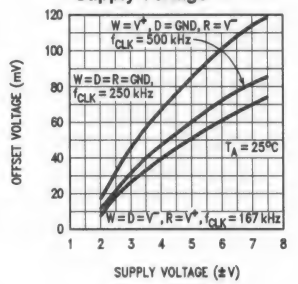
Power Supply Current vs Temperature



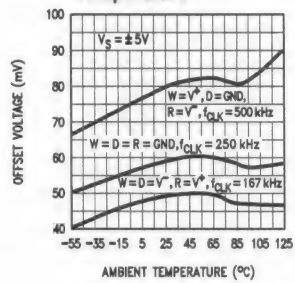
Offset Voltage vs Clock Frequency



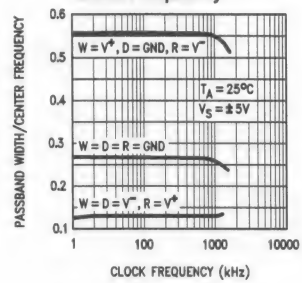
Offset Voltage vs Supply Voltage



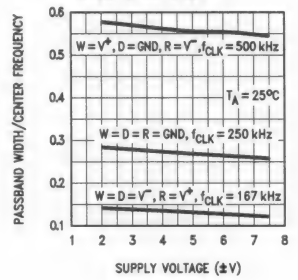
Offset Voltage vs Temperature



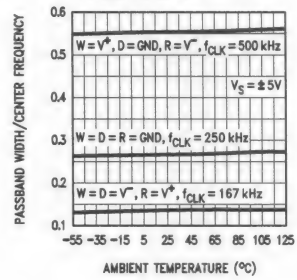
Passband Width vs Clock Frequency



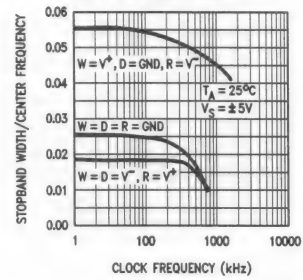
Passband Width vs Supply Voltage



Passband Width vs Temperature


W = V⁺, D = GND, R = V⁻, f_{CLK} = 500 kHz
$V_S = \pm 5V$
W = D = R = GND, f_{CLK} = 250 kHz
W = D = V⁻, R = V⁺, f_{CLK} = 167 kHz
PASSBAND WIDTH/CENTER FREQUENCY
AMBIENT TEMPERATURE (°C)

Stopband Width vs Clock Frequency



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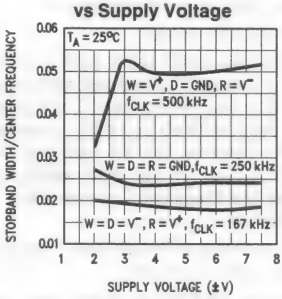
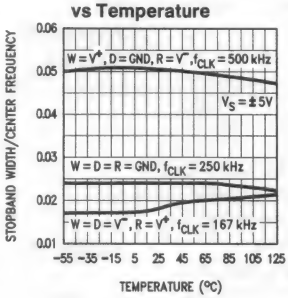
Typical Performance Characteristics (Continued)

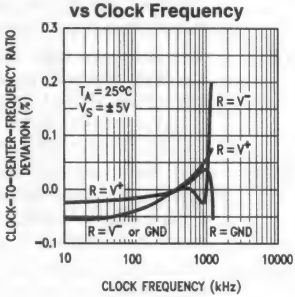
Stopband Width vs Supply Voltage



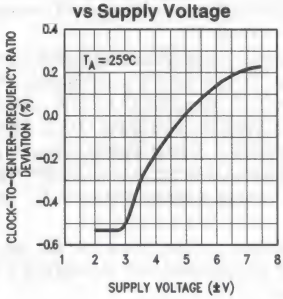
Stopband Width vs Temperature



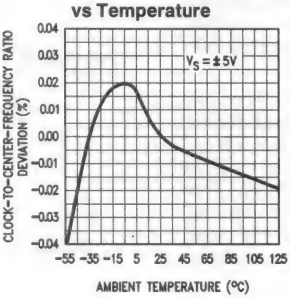
Clock-to-Center-Frequency Ratio Deviation vs Clock Frequency



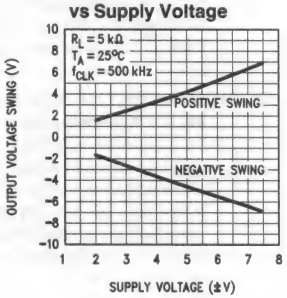
Clock-to-Center-Frequency Ratio Deviation vs Supply Voltage



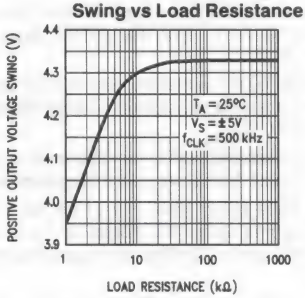
Clock-to-Center-Frequency Ratio Deviation vs Temperature



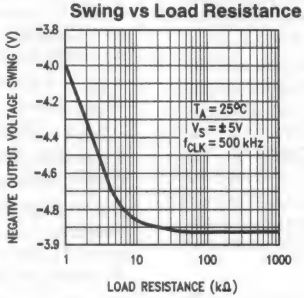
Output Swing vs Supply Voltage



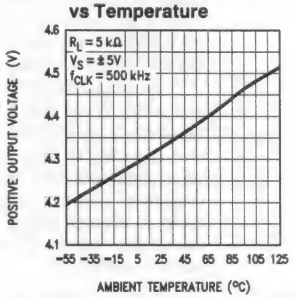
Positive Output Voltage Swing vs Load Resistance



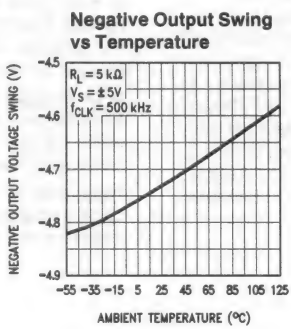
Negative Output Voltage Swing vs Load Resistance



Positive Output Swing vs Temperature



Negative Output Swing vs Temperature



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Pin Descriptions

W (Pin 1) This three-level logic input sets the width of the notch. Notch width is $f_{c2}-f_{c1}$ (see *Figure 1*). When W is tied to V^+ (pin 14), GND (pin 13), or V^- (pin 8), the notch width is 0.55 f_0, 0.26 f_0, or 0.127 f_0, respectively.

R (Pin 2) This three-level logic input sets the ratio of the clock frequency (f_{CLK}) to the center frequency (f_0). When R is tied to V^+, GND, or V^-, the clock-to-center-frequency ratio is 33.33:1, 50:1, or 100:1, respectively.

LD (Pin 3) This three-level logic input sets the division factor of the clock frequency divider. When LD is tied to V^+, GND, or V^-, the division factor is 716, 596, or 2, respectively.

XTAL2 (Pin 4) This is the output of the internal crystal oscillator. When using the internal oscillator, the crystal should be tied between XTAL2 and XTAL1. (The capacitors are internal—no external capacitors are needed for the oscillator to operate.) When not using the internal oscillator this pin should be left open.

XTAL1 (Pin 5) This is the crystal oscillator input. When using the internal oscillator, the crystal should be tied between XTAL1 and XTAL2. XTAL1 can also be used as an input for an external clock signal swinging from V^+ to V^-. The frequency of the crystal or the external clock will be divided internally by the clock divider as determined by the programming voltage on pin 3.

CLK (Pin 6) This is the filter clock pin. The clock signal appearing on this pin is the filter clock (f_{CLK}). When using the internal crystal oscillator or an external clock signal applied to pin 5 while pin 7 is tied to V^+, the CLK pin is the output of the divider and can be used to drive other LMF90s with its rail-to-rail output swing. When not using the internal crystal oscillator or an external clock on pin 5, the CLK pin can be used as a CMOS or TTL clock input provided that pin 7 is tied to GND or V^-. For best performance, the duty cycle of a clock signal applied to this pin should be near 50%, especially at higher clock frequencies.

XLS (Pin 7) This is a three-level logic pin. When XLS is tied to V^+, the crystal oscillator and frequency divider are enabled and CLK (pin 6) is an output. When XLS is tied to GND (pin 13), the crystal oscillator and frequency divider are disabled and pin 6 is an input for a clock swinging between V^- and V^+. When XLS is tied to V^-, the crystal oscillator and frequency divider are disabled and pin 6 is a TTL level clock input for a clock signal swinging between GND and V^+ or between V^- and GND.

V^- (Pin 8) This is the negative power supply pin. It should be bypassed with at least a 0.1 μF capacitor. For single-supply operation, connect this pin to system ground.

V_{OUT} (Pin 9) This is the filter output.

D (Pin 10) This two-level logic input is used to set the depth of the notch (the attenuation at f_0). When D is tied to GND or V^-, the typical notch depth is 48 dB or 39 dB, respectively. Note, however, that the notch depth is also dependent on the width setting (pin 1). See the Electrical Characteristics for tested limits.

V_{IN2} (Pin 11) This is the input to the difference amplifier section of the notch filter.

V_{IN1} (Pin 12) This is the input to the internal bandpass filter. This pin is normally connected to pin 11. For wide bandwidth applications, an anti-aliasing filter can be inserted between pin 11 and pin 12.

GND (Pin 13) This is the analog ground reference for the LMF90. In split supply applications, GND should be connected to the system ground. When operating the LMF90 from a single positive power supply voltage, pin 13 should be connected to a "clean" reference voltage midway between V^+ and V^-.

V^+ (Pin 14) This is the positive power supply pin. It should be bypassed with at least a 0.1 μF capacitor.

1.0 Definition of Terms

A_{max}: the maximum amount of gain variation within the filter's passband (See *Figure 1*). For the LMF90, A_{Max} is nominally equal to 0.25 dB.

A_{min}: the minimum attenuation within the notch's stopband. (See *Figure 1*). This parameter is adjusted by programming voltage applied to pin 10 (D).

Bandwidth (BW) or Passband Width: the difference in frequency between the notch filter's two cutoff frequencies.

Cutoff Frequency: for a notch filter, one of the two frequencies, f_{C1} and f_{C2} that define the edges of the passband. At these two frequencies, the filter has a gain equal to the passband gain.

f_{CLK}: the frequency of the clock signal that appears at the CLK pin. This frequency determines the filter's center frequency. Depending on the programming voltage on pin 2 (R), f_{CLK} will be either 33.33, 50, or 100 times the center frequency of the notch.

f_0 or f_{Notch}: the center frequency of the notch filter. This frequency is measured by finding the two frequencies for which the gain −3 dB relative to the passband gain, and calculating their geometrical mean.

Passband: for a notch filter, frequencies above the upper cutoff frequency (f_{C2} in *Figure 1*) and below the lower cutoff frequency (f_{C1} in *Figure 1*).

1.0 Definition of Terms (Continued)

Passband Gain: the notch filter's gain for signal frequencies near dc or $f_{CLK}/2$. The passband gain of a notch filter is also called "H_{ON}". For the LMF90, the passband gain is nominally 0 dB.

Passband Ripple: the variation in gain within the filter's passband.

Stopband: for a notch filter, the range of frequencies for which the attenuation is at least A_{min} (f_{S1} to f_{S2}) in *Figure 1*).

Stop Frequency: one of the two frequencies (f_{S1} and f_{S2}) at the edges of the notch's stopband.

Stopband Width (SBW): the difference in frequency between the two stopband edges ($f_{S2}-f_{S1}$).

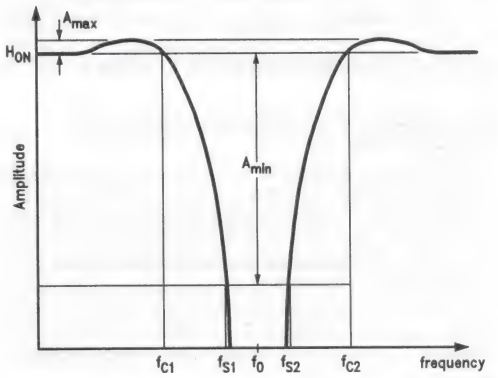

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FIGURE 1. General Form of Notch Response

2.0 Applications Information

2.1 FUNCTIONAL DESCRIPTION

The LMF90 uses switched-capacitor techniques to realize a fourth-order elliptic notch transfer function with 0.25 dB passband ripple. No external components other than supply bypass capacitors and a clock (or crystal) are required.

As is evident from the block diagram, the analog signal path consists of a fourth-order bandpass filter and a summing amplifier. The analog input signal is applied to the input of the bandpass filter, and to one of the summing amplifier inputs. The bandpass filter's output drives the other summing amplifier input. The output of the summing amplifier is the difference between the input signal and the bandpass output, and has a notch filter characteristic. Notch width and depth are controlled by the dc programming voltages applied to two pins (1 and 10), and the center frequency is proportional to the clock frequency, which may be generated externally or internally with the aid of an external crystal. The clock-to-center-frequency ratio can be one of three different values, and is selected by the voltage on a three-level logic input (pin 2).

The clock signal passes through a digital frequency divider circuit that can divide the clock frequency by any of three different factors before it reaches the filters. This divider can also be disabled, if desired. Pin 7 enables and disables the frequency divider and also configures the clock inputs for operation with an external CMOS or TTL clock or with the internal oscillator circuit.

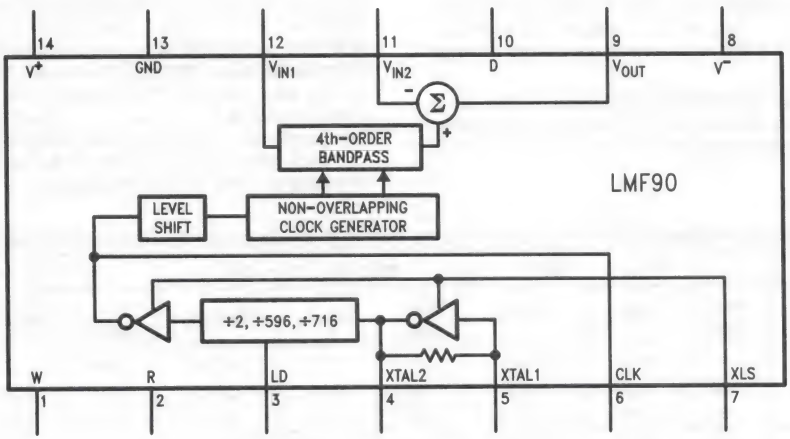

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FIGURE 2. LMF90 Block Diagram

2.0 Applications Information (Continued)

2.2 PROGRAMMING PINS

The LMF90 has five control pins that are used to program the filter's characteristics via a three-level logic scheme. In dual-supply applications, these inputs are tied to either V^+, V^-, or GND in order to select a particular set of characteristics. For example, the W input (pin 1) sets the filter's passband width to 0.55 f_0, 0.26 f_0 or 0.127 f_0 when the W input is connected to V^+, GND, or V^-, respectively. Applying V^- and GND to the D input (pin 10) will set the notch depth to 40 dB or 30 dB, respectively.

The R input (pin 2) is another three-level logic input, and it sets the clock-to-center-frequency ratio to 33.33:1, 50:1, or 100:1 for input voltages equal to V^+, GND, or V^-, respectively. Note that the clock frequency referred to here is the frequency at the CLK pin and at the frequency divider output (if used). This is different from the frequency at the divider's input. LD (pin 3) sets the frequency divider's division factor to either 716, 596, or 2 for input voltages equal to V^+, GND, or V^-, respectively. XLS (pin 7) enables and disables the crystal oscillator and clock divider. When XLS is connected to the positive supply, the oscillator and divider are enabled, and CLK is the output of the divider and can drive the clock inputs of other LMF90s. When XLS is connected to GND, the oscillator and divider are disabled, and the CLK pin becomes a clock input for CMOS-level signals. Connecting XLS to the negative supply disables the oscillator and divider and causes CLK to operate as a TTL-level clock input.

Using an external 3.579545 MHz color television crystal with the internal oscillator and divider, it is possible to build a power line frequency notch for 50 Hz or 60 Hz line frequencies or their second and third harmonics using the LMF90. A 60 Hz notch is shown in the Typical Application circuit on the first page of this data sheet. Connecting LD to V^+ changes the notch frequency to 50 Hz. Changing the clock-to-center-frequency ratio to 50:1 results in a second-harmonic notch, and a 33:1 ratio causes the LMF90 to notch the third harmonic.

Table I illustrates 18 different combinations of filter bandwidth, depth, and clock-to-center-frequency ratio obtained by choosing the appropriate W, D, and R programming voltages.

2.3 DIGITAL INPUTS AND OUTPUTS

As mentioned above, the CLK pin can serve as either an input or an output, depending on the programming voltage on XLS. When CLK is operating as a TTL input, it will operate properly in both dual-supply and single-supply applications, because it has two logic thresholds—one referred to V^-, and one referred to GND. When operating as an output, CLK swings rail-to-rail (CMOS logic levels).

XTAL1 and XTAL2 are the input and output pins for the internal crystal oscillator. When using the internal oscillator (XLS connected to V^+), the crystal is connected between these two pins. When the internal oscillator is not used, XTAL2 should be left open. XTAL1 can be used as an input for an external CMOS-level clock signal swinging from V^- to V^+. The frequency of the crystal or the external clock applied to XTAL1 will be divided by the internal frequency divider as determined by programming voltage on the LD pin.

2.4 SAMPLED-DATA SYSTEM CONSIDERATIONS OUTPUT STEPS

Because the LMF90 uses switched-capacitor techniques, its performance differs in several ways from non-sampled (continuous) circuits. The analog signal at the input to the internal bandpass filter (pin 12) is sampled during each clock cycle, and, since the output voltage can change only once every clock cycle, the result is a discontinuous output signal. The bandpass output takes the form of a series of voltage "steps", as shown in *Figure 3*. The steps are smaller when the clock frequency is much greater than the signal frequency.

Switched-capacitor techniques are used to set the summing amplifier's gain. Its input and feedback "resistors" are actually made from switches and capacitors. Two sets of these "resistors" are alternated during each clock cycle. Each time these gain-setting components are switched, there will be no feedback connected to the op amp for a short period of time (about 50 ns). This generates very low-amplitude output signals at $f_{CLK} + f_{IN}$, $f_{CLK} - f_{IN}$, $2\,f_{CLK} + f_{IN}$, etc. The amplitude of each of these intermodulation components will typically be at least 70 dB below the input signal amplitude and well beyond the spectrum of interest.

TABLE I. Operation of LMF90 Programming Pins. Values given are for nominal levels of attenuation.

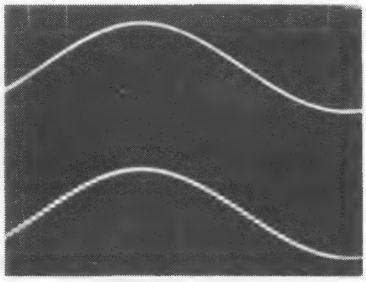
R		V^- (f_{CLK}/f_0 = 100)			GND (f_{CLK}/f_0 = 50)			V^+ (f_{CLK}/f_0 = 33.33)		
D	W	A_{min} (dB)	BW/f_0	SBW/f_0	A_{min} (dB)	BW/f_0	SBW/f_0	A_{min} (dB)	BW/f_0	SBW/f_0
V^-	V^-	−30	0.12	0.019	−30	0.12	0.019	−30	0.12	0.019
	GND	−30	0.26	0.040	−30	0.26	0.040	−30	0.26	0.040
	V^+	−30	0.55	0.082	−30	0.55	0.082	−30	0.55	0.082
GND	V^-	−35	0.12	0.010	−35	0.12	0.010	−35	0.12	0.010
	GND	−40	0.26	0.024	−40	0.26	0.024	−40	0.26	0.024
	V^+	−40	0.55	0.050	−40	0.55	0.050	−40	0.55	0.050

2.0 Applications Information (Continued)

ALIASING

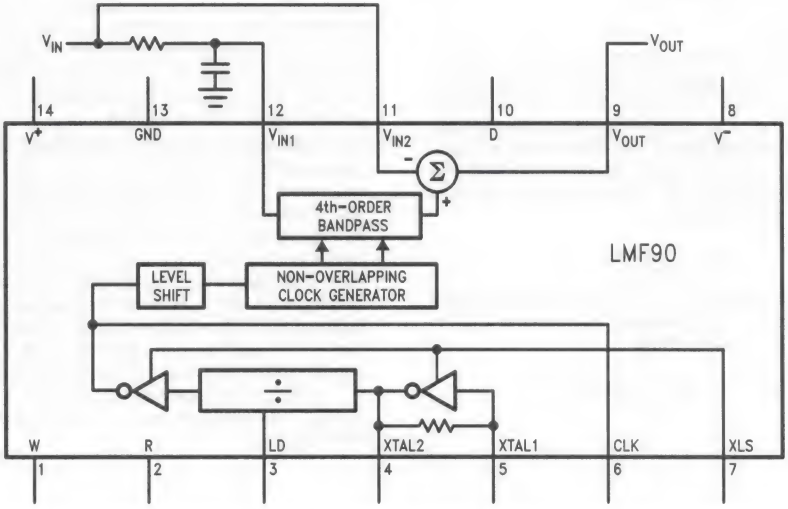
Another important characteristic of sampled-data systems is their effect on signals at frequencies greater than one-half the sampling frequency. (The LMF90's sampling frequency is the same as the filter's clock frequency. This is the frequency at the CLK pin). If a signal with a frequency greater than one-half the sampling frequency is applied to the input of a sampled-data system, it will be "reflected" to a frequency less than one-half the sampling frequency. Thus, an input signal whose frequency is $f_S/2 + 10$ Hz will cause the system to respond as though the input frequency was $f_S/2 - 10$ Hz. This phenomenon is known as "aliasing". Aliasing can be reduced or eliminated by limiting the input signal spectrum to less than $f_S/2$.

In some cases, it may be necessary to use a bandwidth limiting filter (often a simple passive RC low-pass) ahead of the bandpass input. Although the summing amplifier uses switched-capacitor techniques, it does not exhibit aliasing behavior, and the anti-aliasing filter need not be in its input signal path. The filter can be placed ahead of pin 12 as shown in *Figure 4*, with the non-band limited input signal applied to pin 11. The output spectrum will therefore be wideband, although limited by the bandwidth of the summing amplifier's output buffer amplifier (typically 1 MHz), even if f_{CLK} is less than 1 MHz. Phase shift in the anti-aliasing filter will affect the accuracy of the notch transfer function, however, so it is best to use the highest available clock-to-center-frequency ratio (100:1) and set the RC filter cutoff frequency to about 15 to 20 times the notch frequency. This will provide reasonable attenuation of high-frequency input signals, while avoiding degradation of the overall notch response. If the anti-aliasing filter's cutoff frequency is too low, it will introduce phase shift and gain errors large enough to shift the frequency of the notch and reduce its depth. A cutoff frequency that is too high may not provide sufficient attenuation of unwanted high-frequency signals.



TL/H/10354–7

FIGURE 3. Output waveform of a switched-capacitor filter. Note the voltage steps caused by sampling at the clock frequency.



TL/H/10354–8

FIGURE 4. Using a simple passive low-pass filter to prevent aliasing in the presence of high-frequency input signals.

2.0 Applications Information (Continued)

NOISE

Switched-capacitor filters have two kinds of noise at their outputs. There is a random, "thermal" noise component whose level is typically on the order of hundreds of microvolts. The other kind of noise is digital clock feedthrough. This will have an amplitude in the vicinity of 50 mV peak-to-peak. In some applications, the clock noise frequency is so high compared to the signal frequency that it is unimportant. In other cases, clock noise may have to be removed from the output signal with, for example, a passive low-pass filter at the LMF90's output pin.

CLOCK FREQUENCY LIMITATIONS

The performance characteristics of a switched-capacitor filter depend on the switching (clock) frequency. At very low clock frequencies (below 10 Hz), the time between clock cycles is relatively long, and small parasitic leakage currents cause the internal capacitors to discharge sufficiently to affect the filter's offset voltage and gain. This effect becomes more pronounced at elevated operating temperatures.

At higher clock frequencies, performance deviations are primarily due to the reduced time available for the internal operational amplifiers to settle. Best performance with high clock frequencies will be obtained when the filter clock's duty cycle is 50%. The clock frequency divider, when used, provides a 50% duty cycle clock to the filter, but when an external clock is applied to CLK, it should have a duty cycle close to 50% for best performance.

Input Impedance

The input to the bandpass section of the LMF90 (V_{IN1}) is similar to the switched-capacitor circuit shown in *Figure 5*. During the first half of a clock cycle, the θ_1 switch closes, charging C_{IN} to the input voltage V_{IN}. During the second half-cycle, the θ_2 switch closes, and the charge on C_{IN} is transferred to the feedback capacitor. At frequencies well below the clock frequency, the input impedance approximates a resistor whose value is

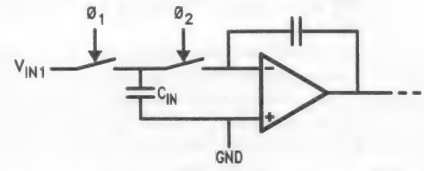
$$R_{IN} = \frac{1}{C_{IN}\, f_{CLK}}.$$

At the bandpass filter input, C_{IN} is nominally 3.0 pF. For a worst-case calculation of effective R_{IN}, assume C_{IN} = 3.0 pF and f_{CLK} = 1.5 MHz. Thus,

$$R_{IN}\,(\text{Min}) = \frac{1}{4.5 \times 10^{-6}} = 222\ \text{k}\Omega.$$

At the maximum clock frequency of 1.5 MHz, the lowest typical value for the effective R_{IN} at the V_{IN1} input is therefore 222 kΩ. Note that R_{IN} increases as f_{CLK} decreases, so the input impedance will be greater than or equal to this value. Source impedance should be low enough that this input impedance doesn't significantly affect gain.

The summing amplifier input impedance at V_{IN2} is calculated in a similar manner, except that C_{IN} = 5.0 pF. This yields a minimum input impedance of 133 kΩ at V_{IN2}. When both inputs are connected together, the combined input impedance will be 83.3 kΩ with a 1.5 MHz filter clock.



TL/H/10354–9

FIGURE 5. Simplified LMF90 bandpass section input stage. At frequencies well below the center frequency, the input impedance appears to be resistive.

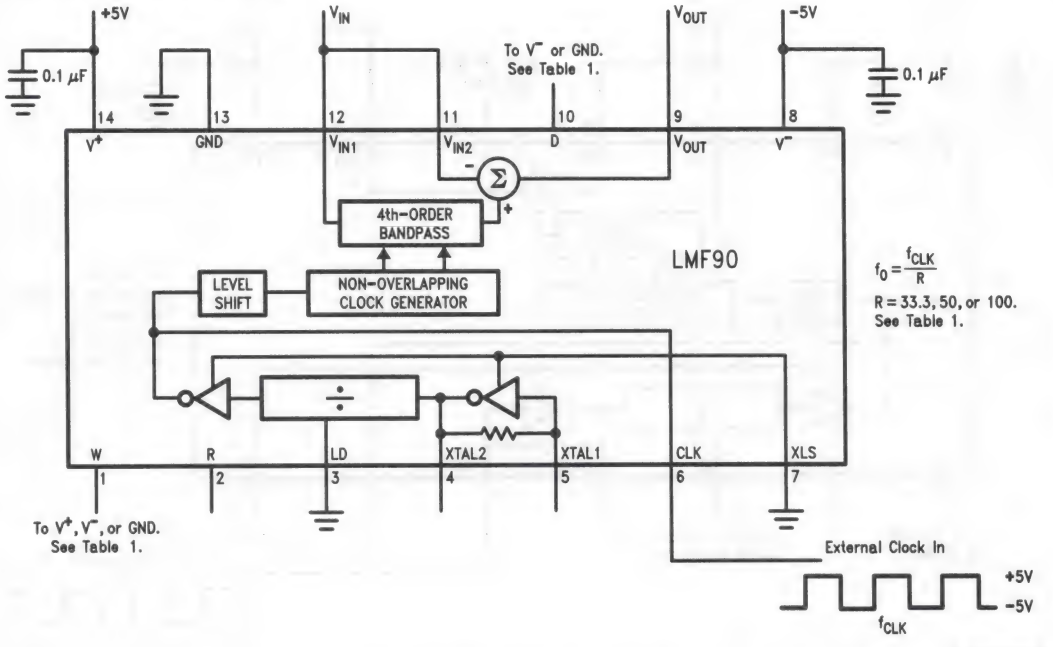
2.5 POWER SUPPLY AND CLOCK OPTIONS

The LMF90 is designed to operate from either single or dual power supply voltages from 5V to 15V. In either case, the supply pins should be well-bypassed to minimize any feedthrough of power supply noise into the filter's signal path. Such feedthrough can significantly reduce the depth of the notch. For operation from dual supply voltages, connect V^- (pin 8) to the negative supply, GND (pin 13) to the system ground, and V^+ to the positive supply.

For single supply operation, simply connect V^- to system ground and GND (Pin 13) to a "clean" reference voltage at mid-supply. This reference voltage can be developed with a pair of resistors and a capacitor as shown in *Figures 10* through *16*. Note that for single supply operation, the three-level logic inputs should be connected to system ground and $V^+/2$ instead of V^- and GND. The CLK input will operate properly with TTL-level clock signals when the LMF90 is powered from either single or dual supplies because it has two TTL thresholds, one referred to the V^- pin and one referred to the GND pin. XLS should be connected to the V^- pin when an external TTL clock is used. *Figures 6* through *16* illustrate a wide variety of power supply and clock options.

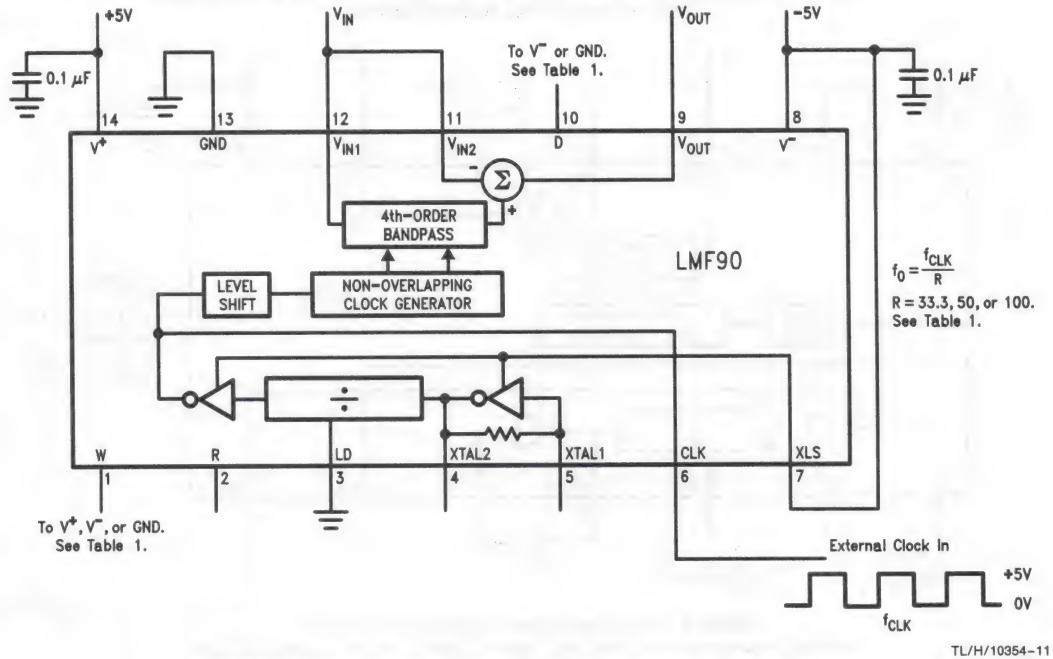
2.0 Applications Information (Continued)

DUAL-SUPPLY CLOCK OPTIONS



TL/H/10354–10

FIGURE 6. Dual supply; external CMOS-level clock. Internal frequency divider disabled.

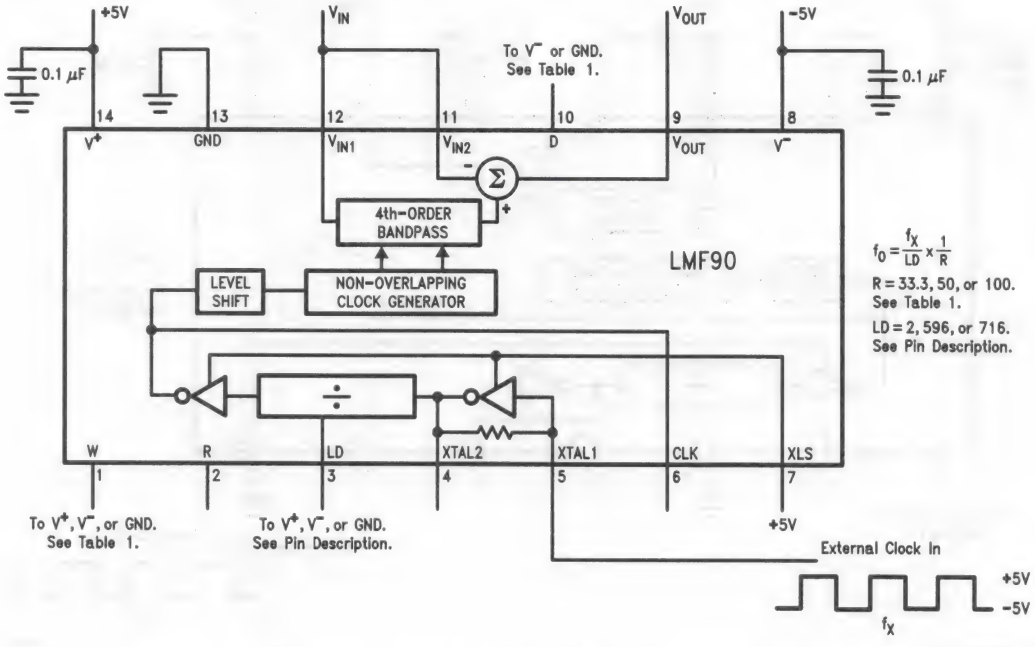


TL/H/10354–11

FIGURE 7. Dual supply; TTL-level clock. Internal frequency divider disabled.

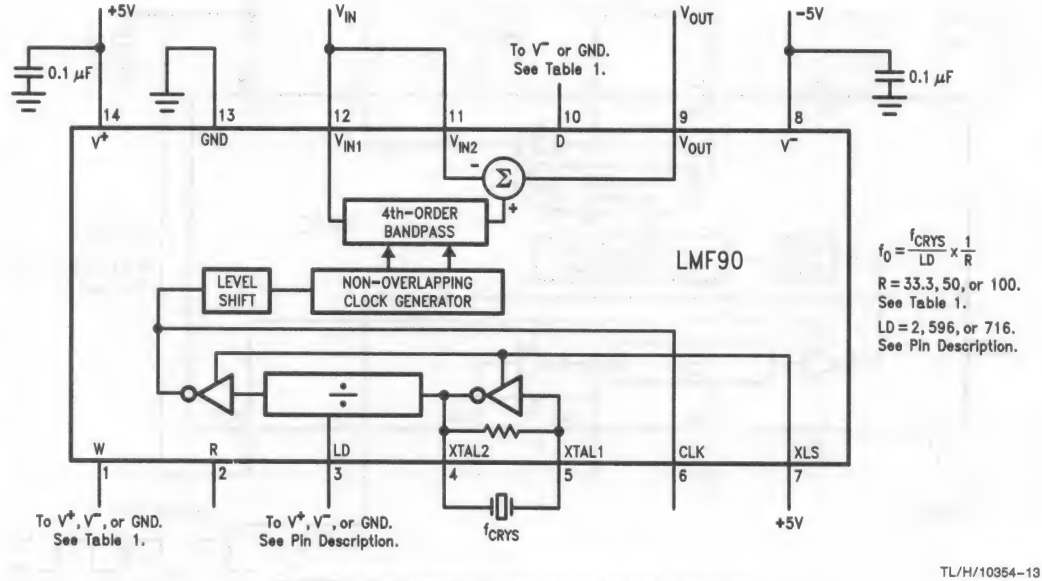
2.0 Applications Information (Continued)

DUAL-SUPPLY CLOCK OPTIONS



TL/H/10354–12

FIGURE 8. Dual Supply; external CMOS-level clock. Internal frequency divider enabled. Output of logic divider available on pin 6.

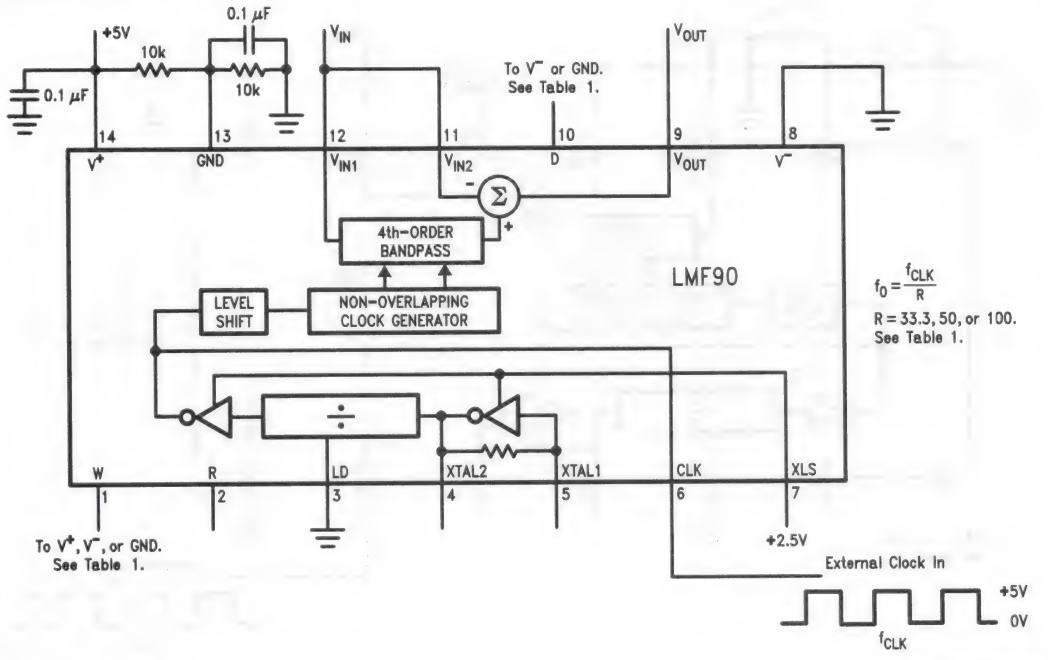


TL/H/10354–13

FIGURE 9. Dual supply; internal crystal clock oscillator. Internal frequency divider enabled. Output of logic divider available on pin 6.

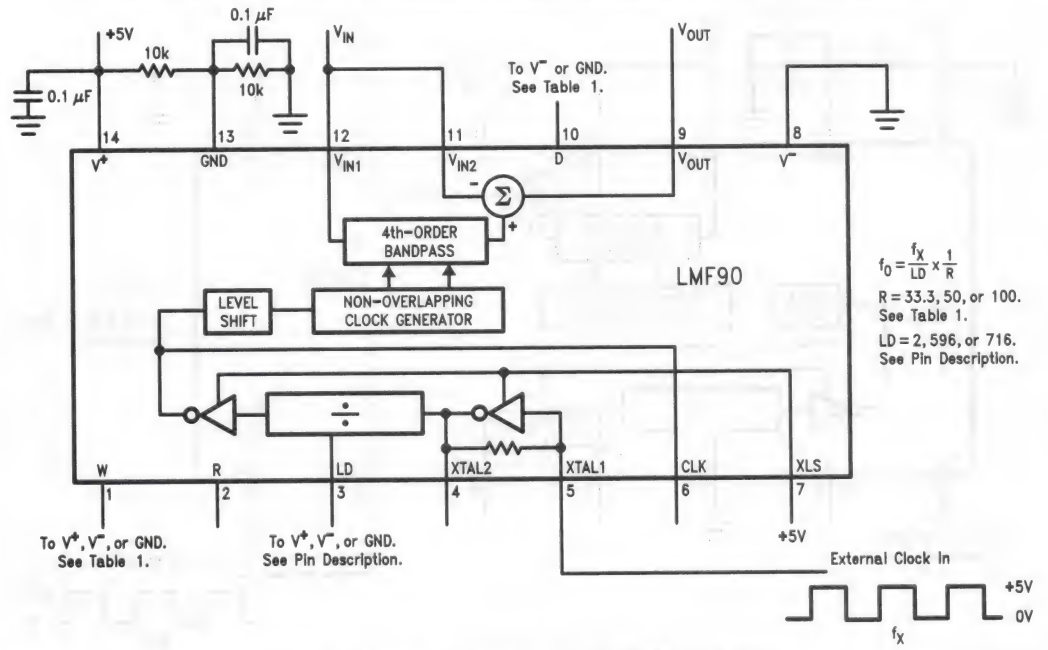
2.0 Applications Information (Continued)

SINGLE-SUPPLY CLOCK OPTIONS



TL/H/10354–14

FIGURE 10. Single +5V supply; external TTL-level clock. Internal frequency divider disabled.

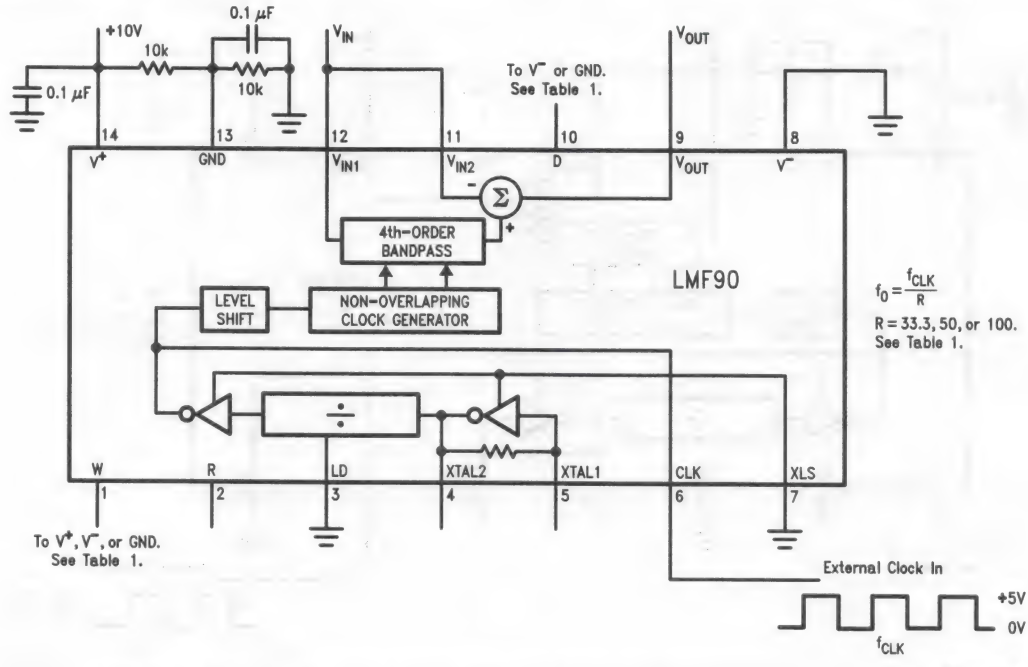


TL/H/10354–15

FIGURE 11. Single +5V supply; external CMOS-level clock. Internal frequency divider enabled. Output of logic divider available on pin 6.

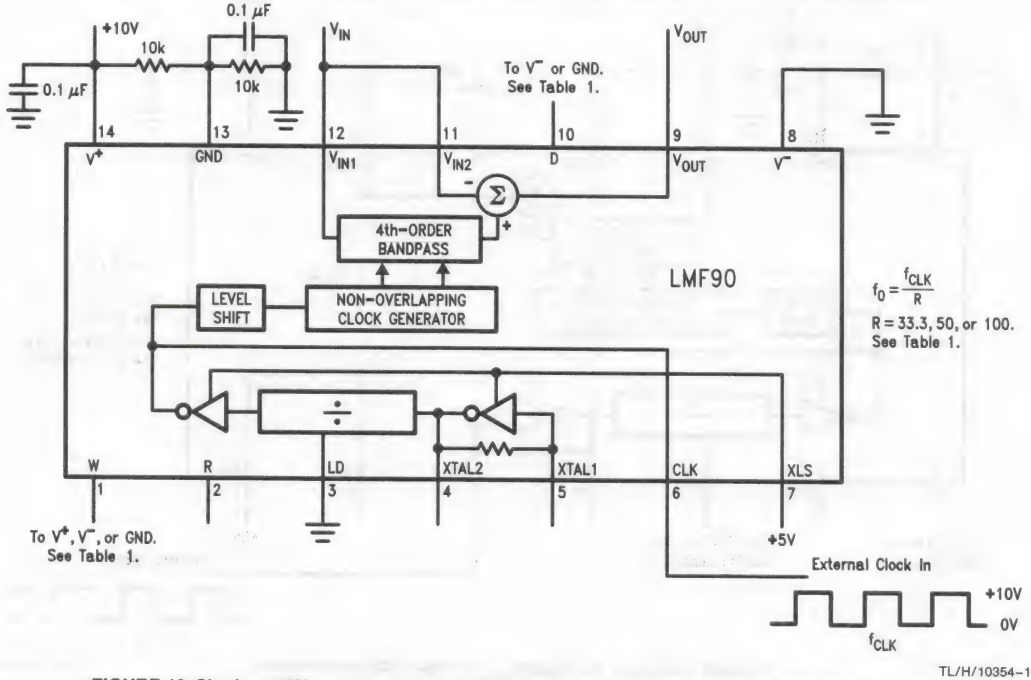
2.0 Applications Information (Continued)

SINGLE-SUPPLY CLOCK OPTIONS



TL/H/10354–16

FIGURE 12. Single + 10V supply; external TTL-level clock. Internal frequency divider disabled.

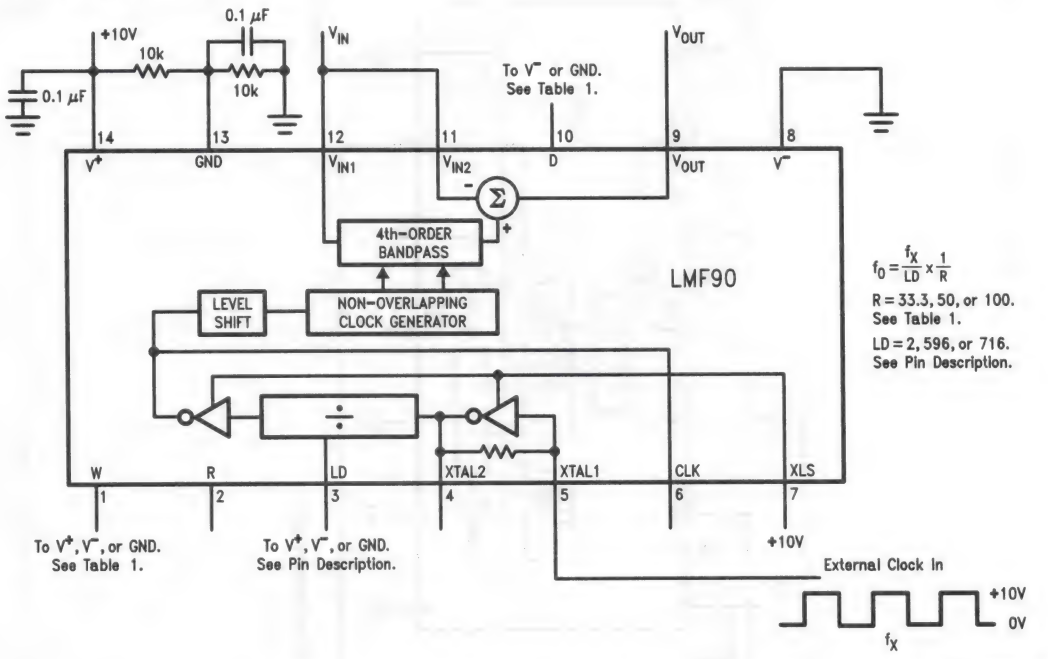


TL/H/10354–17

FIGURE 13. Single + 10V supply; external CMOS-level clock. Internal frequency divider disabled.

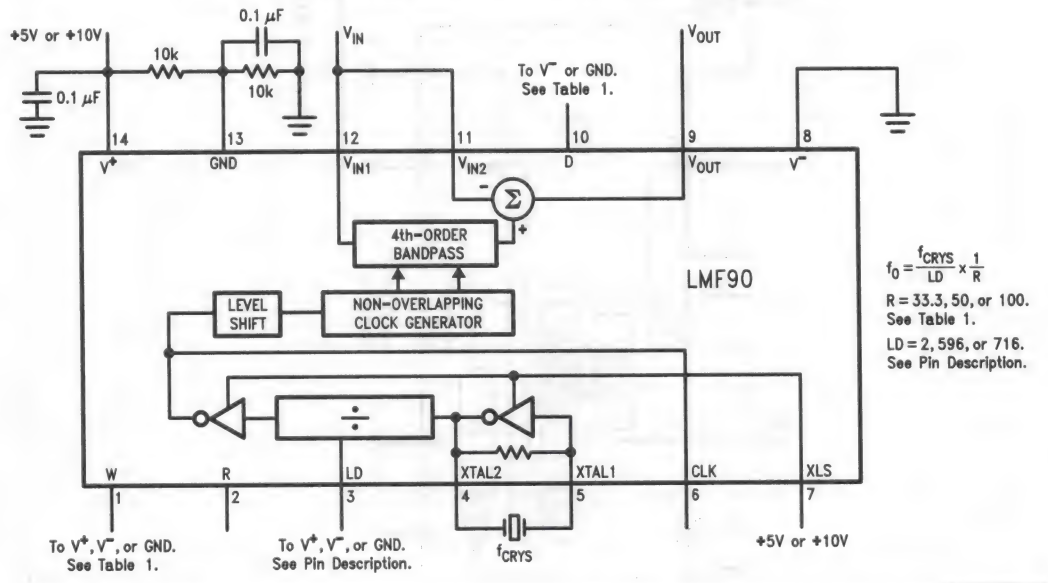
2.0 Applications Information (Continued)

SINGLE-SUPPLY CLOCK OPTIONS



TL/H/10354–18

FIGURE 14. Single + 10V supply; external CMOS-level clock. Internal frequency divider enabled. Output of logic divider available on pin 6.



TL/H/10354–19

FIGURE 15. Single + 5V or + 10V supply; internal crystal clock oscillator. Internal frequency divider enabled. Output of logic divider available on pin 6.

Typical Application

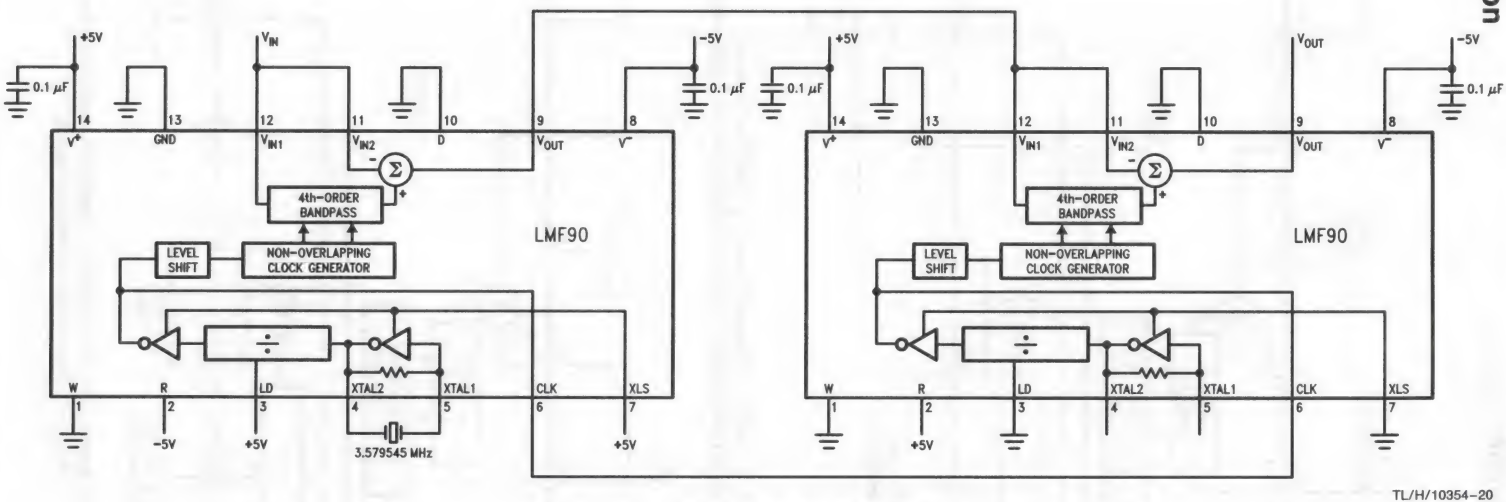


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FIGURE 16. 50 Hz and 150 Hz Notch Filter

National Semiconductor

LMF100 High Performance Dual Switched Capacitor Filter

General Description

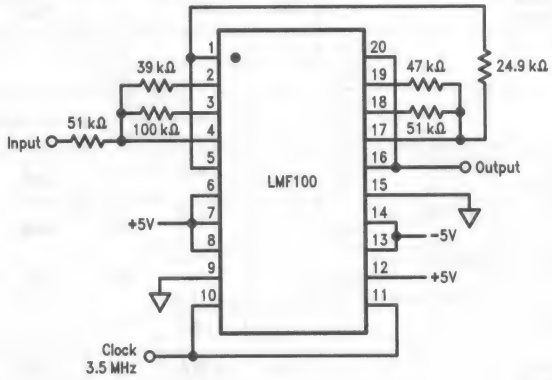
The LMF100 consists of two independent general purpose high performance switched capacitor filters. With an external clock and 2 to 4 resistors, various second-order and first-order filtering functions can be realized by each filter block. Each block has 3 outputs. One output can be configured to perform either an allpass, highpass, or notch function. The other two outputs perform bandpass and lowpass functions. The center frequency of each filter stage is tuned by using an external clock or a combination of a clock and resistor ratio. Up to a 4th-order biquadratic function can be realized with a single LMF100. Higher order filters are implemented by simply cascading additional packages, and all the classical filters (such as Butterworth, Bessel, Elliptic, and Chebyshev) can be realized.

The LMF100 is fabricated on National Semiconductor's high performance analog silicon gate CMOS process, LMCMOS™. This allows for the production of a very low offset, high frequency filter building block. The LMF100 is pin-compatible with the industry standard MF10, but provides greatly improved performance.

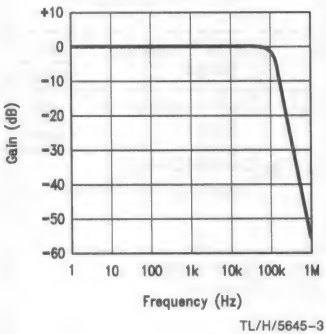
Features

- Wide 4V to 15V power supply range
- Operation up to 100 kHz
- Low offset voltage (50:1 or 100:1 mode) — typically Vos1 = ±5 mV, Vos2 = ±15 mV, Vos3 = ±15 mV
- Low crosstalk −60 dB
- Clock to center frequency ratio accuracy ±0.2% typical
- $f_0 \times Q$ range up to 1.8 MHz
- Pin-compatible with MF10

4th Order 100 kHz Butterworth Lowpass Filter



TL/H/5645–2



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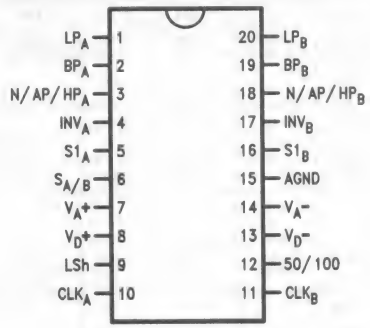
Connection Diagram

Surface Mount and Dual-In-Line Package



TL/H/5645–18

Top View

Order Number LMF100AJ, LMF100CCJ, LMF100ACN, LMF100CCN or LMF100CCWM
See NS Package Number J20A, N20A or M20B

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage ($V^+ - V^-$)	16V
Voltage at Any Pin	$V^+ + 0.3V$ $V^- - 0.3V$
Input Current at Any Pin (Note 2)	5 mA
Package Input Current (Note 2)	20 mA
Power Dissipation (Note 3)	500 mW
Storage Temperature	150°C
ESD Susceptability (Note 11)	2000V

Soldering Information

N Package: 10 sec.	260°C
J Package: 10 sec.	300°C
SO Package: Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (Appendix D) for other methods of soldering surface mount devices.

Operating Ratings (Note 1)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
LMF100ACN, LMF100CCN	$0°C \le T_A \le +70°C$
LMF100CCWM	$0°C \le T_A \le +70°C$
LMF100CCJ	$-40°C \le T_A \le +85°C$
LMF100AJ	$-55°C \le T_A \le +125°C$
Supply Voltage	$4V \le V^+ - V^- \le 15V$

Electrical Characteristics

The following specifications apply for Mode 1, Q = 10 ($R_1 = R_3 = 100k$, $R_2 = 10k$), $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX}**; all other limits $T_A = T_J = 25°C$.

Symbol	Parameter		Conditions		LMF100ACN, LMF100CCN, LMF100CCWM			LMF100AJ LMF100CCJ			Units
					Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
I_s	Maximum Supply Current		f_{CLK} = 250 kHz No Input Signal		9	13	**13**	9	**13**		mA
f_0	Center Frequency Range	MIN			0.1			0.1			Hz
		MAX			100			100			kHz
f_{CLK}	Clock Frequency Range	MIN			5.0			5.0			Hz
		MAX			3.5			3.5			MHz
f_{CLK}/f_0	Clock to Center Frequency Ratio Deviation		V_{Pin12} = 5V or 0V f_{CLK} = 1 MHz	LMF100A	±0.2	±0.6	**±0.6**	±0.2	**±0.6**		%
				LMF100C	±0.2	±0.8	**±0.8**	±0.2	**±0.8**		%
$\frac{\Delta Q}{Q}$	Q Error (MAX) (Note 4)		Q = 10, Mode 1 V_{Pin12} = 5V or 0V f_{CLK} = 1 MHz	LMF100A	±0.5	±4	**±5**	±0.5	**±5**		%
				LMF100C	±0.5	±5	**±6**	±0.5	**±6**		%
H_{OBP}	Bandpass Gain at f_0		f_{CLK} = 1 MHz		0	±0.4	**±0.4**	0	**±0.4**		dB
H_{OLP}	DC Lowpass Gain		$R_1 = R_2$ = 10k f_{CLK} = 250 kHz		0	±0.2	**±0.2**	0	**±0.2**		dB
V_{OS1}	DC Offset Voltage (Note 5)		f_{CLK} = 250 kHz		±5.0	±15	**±15**	±5.0	**±15**		mV
V_{OS2}	DC Offset Voltage (Note 5)		f_{CLK} = 250 kHz	$S_{A/B} = V^+$	±30	±80	**±80**	±30	**±80**		mV
				$S_{A/B} = V^-$	±15	±70	**±70**	±15	**±70**		mV
V_{OS3}	DC Offset Voltage (Note 5)		f_{CLK} = 250 kHz		±15	±40	**±60**	±15	**±60**		mV
	Crosstalk (Note 6)		A Side to B Side or B Side to A Side		−60			−60			dB
	Output Noise (Note 12)		f_{CLK} = 250 kHz 20 kHz Bandwidth 100:1 Mode	N	40			40			μV
				BP	320			320			
				LP	300			300			
	Clock Feedthrough (Note 13)		f_{CLK} = 250 kHz 100:1 Mode		6			6			mV
V_{OUT}	Minimum Output Voltage Swing		R_L = 5k (All Outputs)		+4.0 −4.7	±3.8	**±3.7**	+4.0 −4.7	**±3.7**		V
			R_L = 3.5k (All Outputs)		+3.9 −4.6			+3.9 −4.6			V
GBW	Op Amp Gain BW Product				5			5			MHz
SR	Op Amp Slew Rate				20			20			V/μs

Electrical Characteristics

The following specifications apply for Mode 1, Q = 10 ($R_1 = R_3 = 100k$, $R_2 = 10k$), $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX}**; all other limits $T_A = T_J = 25°C$. (Continued)

Symbol	Parameter		Conditions	LMF100ACN, LMF100CCN, LMF100CCWM			LMF100AJ LMF100CCJ			Units
				Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
I_{SC}	Maximum Output Short Circuit Current (Note 7)	Source	(All Outputs)	12			12			mA
		Sink		45			45			mA
I_{IN}	Input Current on Pins: 4, 5, 6, 9, 10, 11, 12, 16, 17				10			**10**		μA

Electrical Characteristics

The following specifications apply for Mode 1, Q = 10 ($R_1 = R_3 = 100k$, $R_2 = 10k$), $V^+ = +2.50V$ and $V^- = -2.50V$ unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX}**; all other limits $T_A = T_J = 25°C$.

Symbol	Parameter		Conditions		LMF100ACN, LMF100CCN, LMF100CCWM			LMF100AJ LMF100CCJ			Units
					Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
I_S	Maximum Supply Current		f_{CLK} = 250 kHz No Input Signal		8	12	**12**	8	**12**		mA
f_0	Center Frequency Range	MIN			0.1			0.1			Hz
		MAX			50			50			kHz
f_{CLK}	Clock Frequency Range	MIN			5.0			5.0			Hz
		MAX			1.5			1.5			MHz
f_{CLK}/f_0	Clock to Center Frequency Ratio Deviation		V_{Pin12} = 2.5V or 0V, f_{CLK} = 1 MHz	LMF100A	±0.2	±0.6	**±0.8**	±0.2	**±0.8**		%
				LMF100C	±0.2	±1	**±1**	±0.2	**±1**		%
$\frac{\Delta Q}{Q}$	Q Error (MAX) (Note 4)		Q = 10, Mode 1, V_{Pin12} = 5V or 0V, f_{CLK} = 1 MHz	LMF100A	±0.5	±4	**±6**	±0.5	**±6**		%
				LMF100C	±0.5	±5	**±8**	±0.5	**±8**		%
H_{OBP}	Bandpass Gain at f_0		f_{CLK} = 1 MHz		0	±0.4	**±0.5**	0	**±0.5**		dB
H_{OLP}	DC Lowpass Gain		$R_1 = R_2$ = 10k, f_{CLK} = 250 kHz		0	±0.2	**±0.2**	0	**±0.2**		dB
V_{OS1}	DC Offset Voltage (Note 5)		f_{CLK} = 250 kHz		±5.0	±15	**±15**	±5.0	**±15**		mV
V_{OS2}	DC Offset Voltage (Note 5)		f_{CLK} = 250 kHz	$S_{A/B} = V^+$	±20	±60	**±60**	±20	**±60**		mV
				$S_{A/B} = V^-$	±10	±50	**±60**	±10	**±60**		mV
V_{OS3}	DC Offset Voltage (Note 5)		f_{CLK} = 250 kHz		±10	±25	**±30**	±10	**±30**		mV
	Crosstalk (Note 6)		A Side to B Side or B Side to A Side		−65			−65			dB
	Output Noise (Note 12)		f_{CLK} = 250 kHz, 20 kHz Bandwidth, 100:1 Mode	N	25			25			μV
				BP	250			250			
				LP	220			220			
	Clock Feedthrough (Note 13)		f_{CLK} = 250 kHz 100:1 Mode		2			2			mV
V_{OUT}	Minimum Output Voltage Swing		R_L = 5k (All Outputs)		+1.6 −2.2	±1.5	**±1.4**	+1.6 −2.2	**±1.4**		V
			R_L = 3.5k (All outputs)		+1.5 −2.1			+1.5 −2.1			V
GBW	Op Amp Gain BW Product				5			5			MHz
SR	Op Amp Slew Rate				18			18			V/μs
I_{SC}	Maximum Output Short Circuit Current (Note 7)	Source	(All Outputs)		10			10			mA
		Sink			20			20			mA

Logic Input Characteristics Boldface limits apply for T_{MIN} to T_{MAX}; all other limits $T_A = T_J = 25°C$.

Parameter		Conditions	LMF100ACN, LMF100CCN, LMF100CCWM			LMF100AJ, LMF100CCJ			Units
			Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
CMOS Clock Input Voltage	MIN Logical "1"	$V^+ = +5V, V^- = -5V,$ $V_{LSh} = 0V$		+3.0	**+3.0**		**+3.0**		V
	MAX Logical "0"			−3.0	**−3.0**		**−3.0**		V
	MIN Logical "1"	$V^+ = +10V, V^- = 0V,$ $V_{LSh} = +5V$		+8.0	**+8.0**		**+8.0**		V
	MAX Logical "0"			+2.0	**+2.0**		**+2.0**		V
TTL Clock Input Voltage	MIN Logical "1"	$V^+ = +5V, V^- = -5V,$ $V_{LSh} = 0V$		+2.0	**+2.0**		**+2.0**		V
	MAX Logical "0"			+0.8	**+0.8**		**+0.8**		V
	MIN Logical "1"	$V^+ = +10V, V^- = 0V,$ $V_{LSh} = 0V$		+2.0	**+2.0**		**+2.0**		V
	MAX Logical "0"			+0.8	**+0.8**		**+0.8**		V
CMOS Clock Input Voltage	MIN Logical "1"	$V^+ = +2.5V, V^- = -2.5V,$ $V_{LSh} = 0V$		+1.5	**+1.5**		**+1.5**		V
	MAX Logical "0"			−1.5	**−1.5**		**−1.5**		V
	MIN Logical "1"	$V^+ = +5V, V^- = 0V,$ $V_{LSh} = +2.5V$		+4.0	**+4.0**		**+4.0**		V
	MAX Logical "0"			+1.0	**+1.0**		**+1.0**		V
TTL Clock Input Voltage	MIN Logical "1"	$V^+ = +5V, V^- = 0V,$ $V_{LSh} = 0V, V_D{}^+ = 0V$		+2.0	**+2.0**		**+2.0**		V
	MAX Logical "0"			+0.8	**+0.8**		**+0.8**		V

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional. These ratings do not guarantee specific performance limits, however. For guaranteed specifications and test conditions, see the Electrical Characteristics. The guaranteed specifications apply only for the test conditions listed. Some performance characteristics may degrade when the device is not operated under the listed test conditions.

Note 2: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 5 mA or less. The sum of the currents at all pins that are driven beyond the power supply voltages should not exceed 20 mA.

Note 3: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 125°C$, and the typical junction-to-ambient thermal resistance of the LMF100ACN/CCN when board mounted is 55°C/W. For the LMF100AJ/CCJ, this number increases to 95°C/W and for the LMF100CCWM this number is 66°C/W.

Note 4: The accuracy of the Q value is a function of the center frequency (f_0). This is illustrated in the curves under the heading "Typical Peformance Characteristics".

Note 5: V_{os1}, V_{os2}, and V_{os3} refer to the internal offsets as discussed in the Applications Information section 3.4.

Note 6: Crosstalk between the internal filter sections is measured by applying a 1 V_{RMS} 10 kHz signal to one bandpass filter section input and grounding the input of the other bandpass filter section. The crosstalk is the ratio between the output of the grounded filter section and the 1 V_{RMS} input signal of the other section.

Note 7: The short circuit source current is measured by forcing the output that is being tested to its maximum positive voltage swing and then shorting that output to the negative supply. The short circuit sink current is measured by forcing the output that is being tested to its maximum negative voltage swing and then shorting that output to the positive supply. These are the worst case conditions.

Note 8: Typicals are at 25°C and represent most likely parametric norm.

Note 9: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 10: Design limits are guaranteed to National's AOQL (Average Outgoing Quality Level) but are not 100% tested.

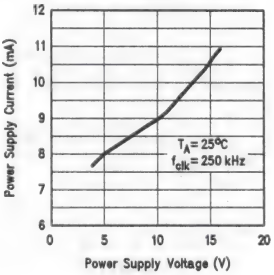
Note 11: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 12: In 50:1 mode the output noise is 3 dB higher.

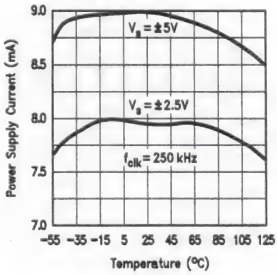
Note 13: In 50:1 mode the clock feedthrough is 6 dB higher.

Typical Performance Characteristics

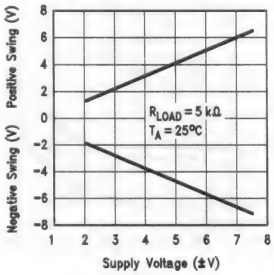
Power Supply Current vs Power Supply Voltage



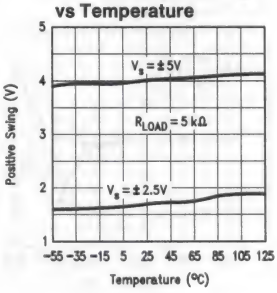
Power Supply Current vs Temperature



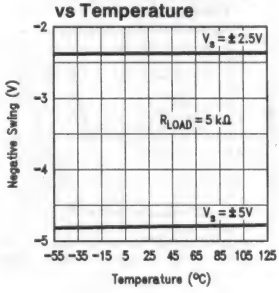
Output Swing vs Supply Voltage



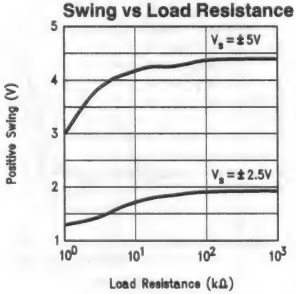
Positive Output Swing vs Temperature



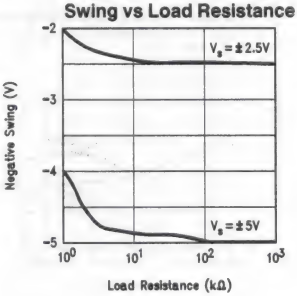
Negative Output Swing vs Temperature



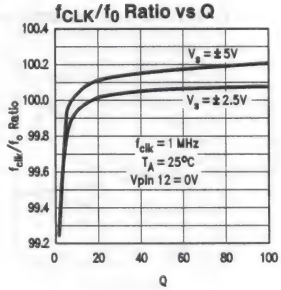
Positive Output Voltage Swing vs Load Resistance



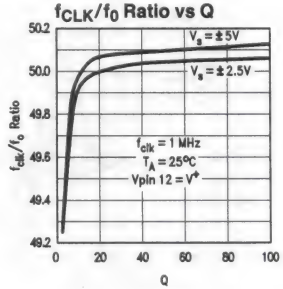
Negative Output Voltage Swing vs Load Resistance



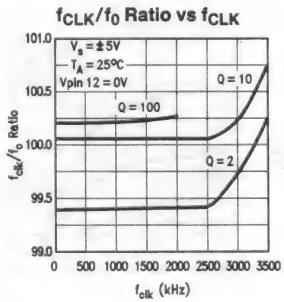
f_{CLK}/f_0 Ratio vs Q



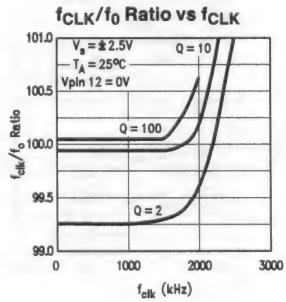
f_{CLK}/f_0 Ratio vs Q



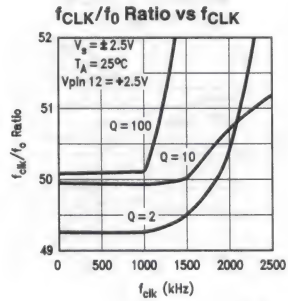
f_{CLK}/f_0 Ratio vs f_{CLK}



f_{CLK}/f_0 Ratio vs f_{CLK}



f_{CLK}/f_0 Ratio vs f_{CLK}



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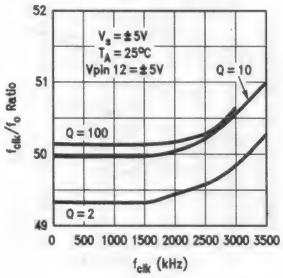
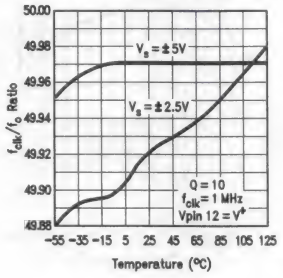
Typical Performance Characteristics (Continued)

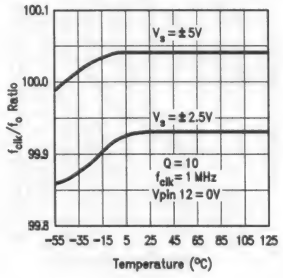
f_{CLK}/f_0 Ratio vs f_{CLK}



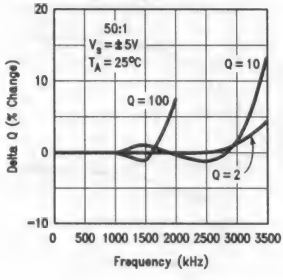
f_{CLK}/f_0 Ratio vs Temperature



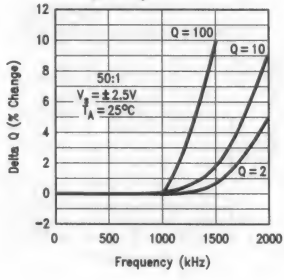
f_{CLK}/f_0 Ratio vs Temperature



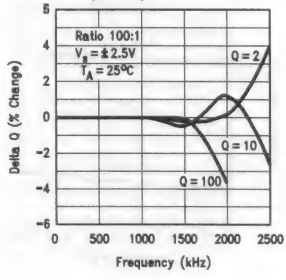
Q Deviation vs Clock Frequency



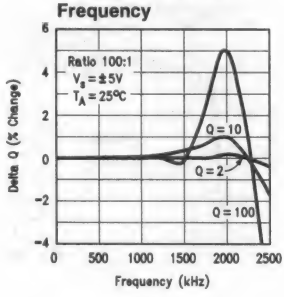
Q Deviation vs Clock Frequency



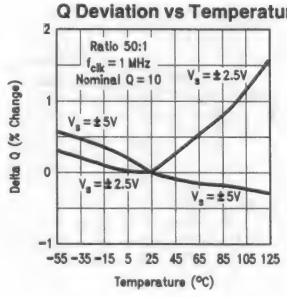
Q Deviation vs Clock Frequency



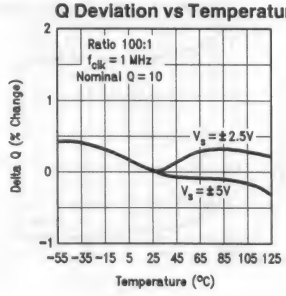
Q Deviation vs Clock Frequency



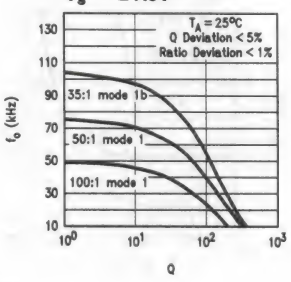
Q Deviation vs Temperature



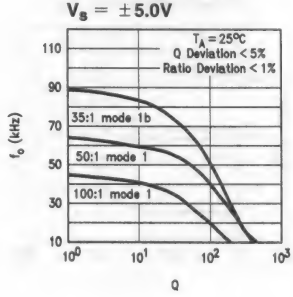
Q Deviation vs Temperature



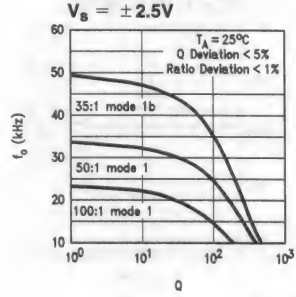
Maximum f_0 vs Q at $V_s = \pm 7.5V$



Maximum f_0 vs Q at $V_s = \pm 5.0V$



Maximum f_0 vs Q at $V_s = \pm 2.5V$



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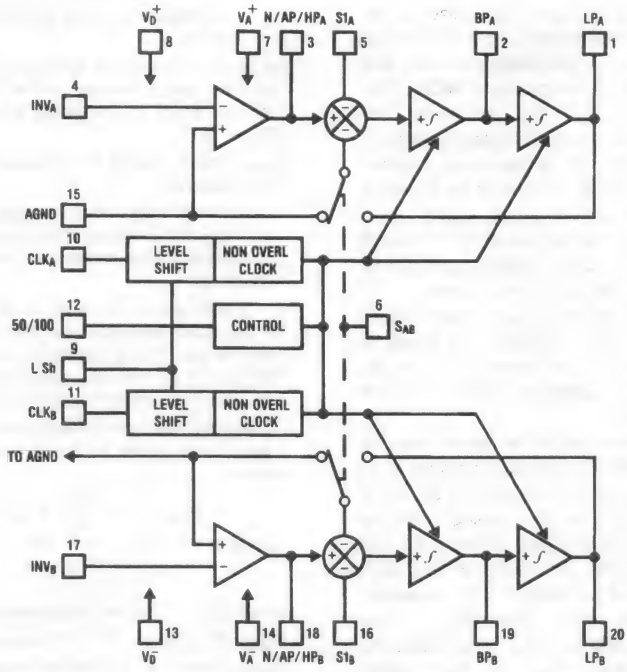
LMF100 System Block Diagram



TL/H/5645–1

Pin Descriptions

LP(1,20), BP(2,19), N/AP/HP(3,18) — The second order lowpass, bandpass and notch/allpass/highpass outputs. These outputs can typically swing to within 1V of each supply when driving a 5 kΩ load. For optimum performance, capacitive loading on these outputs should be minimized. For signal frequencies above 15 kHz the capacitance loading should be kept below 30 pF.

INV(4,17) — The inverting input of the summing opamp of each filter. These are high impedance inputs. The non-inverting input is internally tied to AGND so the opamp can be used only as an inverting amplifier.

S1(5,16) — S1 is a signal input pin used in modes 1b, 4, and 5. The input impedance is $1/f_{CLK} \times 1$ pF. The pin should be driven with a source impedance of less than 1 kΩ. If S1 is not driven with a signal it should be tied to AGND (mid-supply).

$S_{A/B}$(6) — This pin activates a switch that connects one of the inputs of each filter's second summer either to AGND ($S_{A/B}$ tied to V^-) or to the lowpass (LP) output ($S_{A/B}$ tied to V^+). This offers the flexibility needed for configuring the filter in its various modes of operation.

V_A^+(7)* — This is both the analog and digital positive supply.

V_D^+(8)* — This pin needs to be tied to V^+ except when the device is to operate on a single 5V supply and a TTL level clock is applied. For 5V, TTL operation, V_D^+ should be tied to ground (0V).

V_A^-(14), V_D^-(13) — Analog and digital negative supplies. V_A^- and V_D^- should be derived from the same source. They have been brought out separately so they can be bypassed by separate capacitors, if desired. They can also be tied together externally and bypassed with a single capacitor.

Pin Descriptions (Continued)

LSh(9)	Level shift pin. This is used to accommodate various clock levels with dual or single supply operation. With dual ±5V supplies and CMOS (±5V) or TTL (0V–5V) clock levels, LSh should be tied to system ground. For 0V–10V single supply operation the AGND pin should be biased at +5V and the LSh pin should be tied to the system ground for TTL clock levels. LSh should be biased at +5V for ±5V CMOS clock levels. The LSh pin is tied to system ground for ±2.5V operation. For single 5V operation the LSh and V_D+ pins are tied to system ground for TTL clock levels.
CLK(10,11)	Clock inputs for the two switched capacitor filter sections. Unipolar or bipolar clock levels may be applied to the CLK inputs according to the programming voltage applied to the LSh pin. The duty cycle of the clock should be close to 50%, especially when clock frequencies above 200 kHz are used. This allows the maximum time for the internal opamps to settle, which yields optimum filter performance.
50/100(12)*	By tying this pin to V^+ a 50:1 clock to filter center frequency ratio is obtained. Tying this pin at mid-supply (i.e., system ground with dual supplies) or to V^- allows the filter to operate at a 100:1 clock to center frequency ratio.
AGND(15)	This is the analog ground pin. This pin should be connected to the system ground for dual supply operation or biased to mid-supply for single supply operation. For a further discussion of mid-supply biasing techniques see the Applications Information (Section 3.2). For optimum filter performance a "clean" ground must be provided.

*This device is pin-for-pin compatible with the MF10 except for the following changes:

1. Unlike the MF10, the LMF100 has a single positive supply pin (V_A+).
2. On the LMF100 V_D+ is a control pin and is not the digital positive supply as on the MF10.
3. Unlike the MF10, the LMF100 does not support the current limiting mode. When the 50/100 pin is tied to V^- the LMF100 will remain in the 100:1 mode.

1.0 Definitions of Terms

f_{CLK}: the frequency of the external clock signal applied to pin 10 or 11.

f_0: center frequency of the second order function complex pole pair. f_0 is measured at the bandpass outputs of the LMF100, and is the frequency of maximum bandpass gain. *(Figure 1).*

f_{notch}: the frequency of minimum (ideally zero) gain at the notch outputs.

f_z: the center frequency of the second order complex zero pair, if any. If f_z is different from f_0 and if Q_z is high, it can be observed as the frequency of a notch at the allpass output. *(Figure 13).*

Q: "quality factor" of the 2nd order filter. Q is measured at the bandpass outputs of the LMF100 and is equal to f_0 divided by the −3 dB bandwidth of the 2nd order bandpass filter *(Figure 1).* The value of Q determines the shape of the 2nd order filter responses as shown in *Figure 6.*

Q_z: the quality factor of the second order complex zero pair, if any. Q_Z is related to the allpass characteristic, which is written:

$$H_{AP}(s) = \frac{H_{OAP}\left(s^2 - \frac{s\omega_0}{Q_z} + \omega_0^2\right)}{s^2 + \frac{s\omega_0}{Q} + \omega_0^2}$$

where $Q_Z = Q$ for an all-pass response.

H_{OBP}: the gain (in V/V) of the bandpass output at $f = f_0$.

H_{OLP}: the gain (in V/V) of the lowpass output as $f \rightarrow 0$ Hz *(Figure 2).*

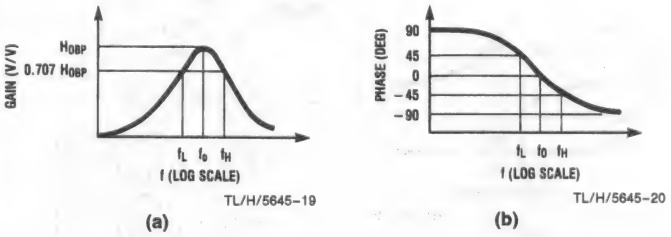
H_{OHP}: the gain (in V/V) of the highpass output as $f \rightarrow f_{CLK}/2$ *(Figure 3).*

H_{ON}: the gain (in V/V) of the notch output as $f \rightarrow 0$ Hz and as $f \rightarrow f_{CLK}/2$, when the notch filter has equal gain above and below the center frequency *(Figure 4).* When the low-frequency gain differs from the high-frequency gain, as in modes 2 and 3a *(Figures 10* and *12),* the two quantities below are used in place of H_{ON}.

H_{ON1}: the gain (in V/V) of the notch output as $f \rightarrow 0$ Hz.

H_{ON2}: the gain (in V/V) of the notch output as $f \rightarrow f_{CLK}/2$.

1.0 Definitions of Terms (Continued)



$$H_{BP}(s) = \frac{H_{OBP}s}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

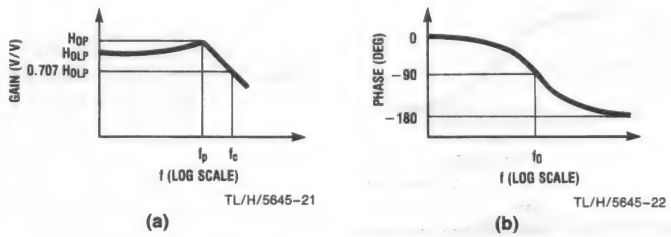
$$Q = \frac{f_0}{f_H - f_L};\ f_0 = \sqrt{f_L f_H}$$

$$f_L = f_0\left(\frac{-1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

$$f_H = f_0\left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

$$\omega_o = 2\pi f_0$$

FIGURE 1. 2nd-Order Bandpass Response



$$H_{LP}(s) = \frac{H_{OLP}\omega_O^2}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

$$f_c = f_0 \times \sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}$$

$$f_p = f_0\sqrt{1 - \frac{1}{2Q^2}}$$

$$H_{OP} = H_{OLP} \times \frac{1}{\frac{1}{Q}\sqrt{1 - \frac{1}{4Q^2}}}$$

FIGURE 2. 2nd-Order Low-Pass Response

$$H_{HP}(s) = \frac{H_{OHP}s^2}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

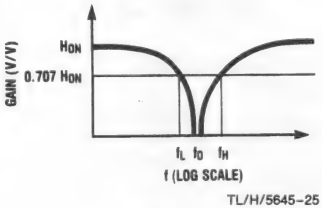
$$f_c = f_0 \times \left[\sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}\right]^{-1}$$

$$f_p = f_0 \times \left[\sqrt{1 - \frac{1}{2Q^2}}\right]^{-1}$$

$$H_{OP} = H_{OHP} \times \frac{1}{\frac{1}{Q}\sqrt{1 - \frac{1}{4Q^2}}}$$

FIGURE 3. 2nd-Order High-Pass Response

1.0 Definitions of Terms (Continued)



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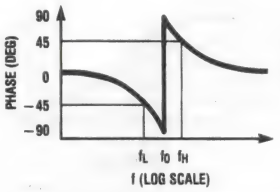
(a)



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(b)

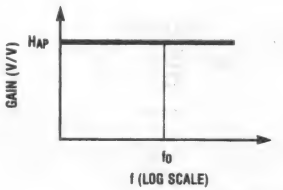
$$H_N(s) = \frac{H_{ON}(s^2 + \omega_o^2)}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

$$Q = \frac{f_0}{f_H - f_L};\ f_0 = \sqrt{f_L f_H}$$

$$f_L = f_0\left(\frac{-1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

$$f_H = f_0\left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

FIGURE 4. 2nd-Order Notch Response



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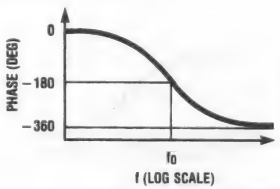
(a)



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(b)

$$H_{AP}(s) = \frac{H_{OAP}\left(s^2 - \frac{s\omega_o}{Q} + \omega_o^2\right)}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

FIGURE 5. 2nd-Order All-Pass Response

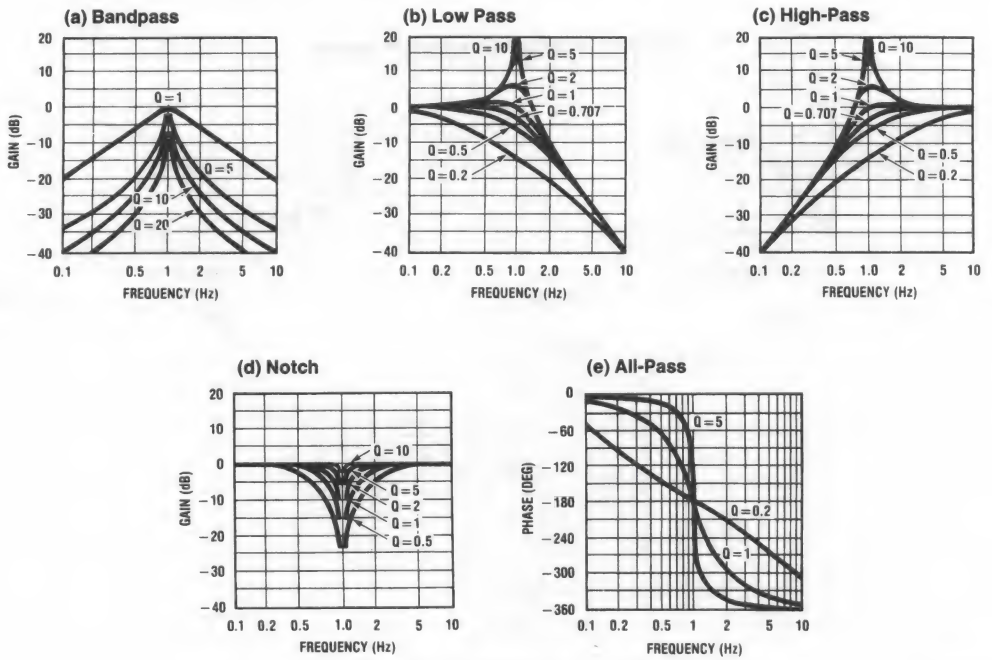

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FIGURE 6. Response of various 2nd-order filters as a function of Q. Gains and center frequencies are normalized to unity.

2.0 Modes of Operation

The LMF100 is a switched capacitor (sampled data) filter. To fully describe its transfer functions, a time domain analysis is appropriate. Since this is cumbersome, and since the LMF100 closely approximates continuous filters, the following discussion is based on the well-known frequency domain. Each LMF100 can produce two full 2nd order functions. See Table I for a summary of the characteristics of the various modes.

MODE 1: Notch 1, Bandpass, Lowpass Outputs:

$f_{notch} = f_0$ **(See *Figure 7*)**

f_0 = center frequency of the complex pole pair

$= \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

f_{notch} = center frequency of the imaginary zero pair = f_0.

H_{OLP} = Lowpass gain (as $f \rightarrow 0$) $= -\frac{R2}{R1}$

H_{OBP} = Bandpass gain (at $f = f_0$) $= -\frac{R3}{R1}$

H_{ON} = Notch output gain as $f \rightarrow 0$ or $f \rightarrow f_{CLK}/2$ $= \frac{-R_2}{R_1}$

$Q = \frac{f_0}{BW} = \frac{R3}{R2}$

= quality factor of the complex pole pair

BW = the −3 dB bandwidth of the bandpass output.

Circuit dynamics:

$H_{OLP} = \frac{H_{OBP}}{Q}$ or $H_{OBP} = H_{OLP} \times Q$

$= H_{ON} \times Q.$

$H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

MODE 1a: Non-Inverting BP, LP (See *Figure 8*)

$f_0 = \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

$Q = \frac{R3}{R2}$

$H_{OLP} = -1$; $H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

$H_{OBP_1} = -\frac{R3}{R2}$

$H_{OBP_2} = 1$ (non-inverting)

Circuit dynamics: $H_{OBP_1} = Q$

Note: V_{IN} should be driven from a low impedance (<1 kΩ) source.

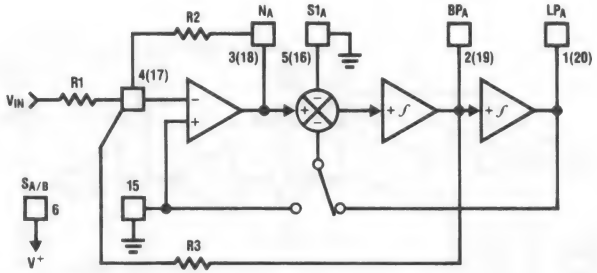

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FIGURE 7. MODE 1

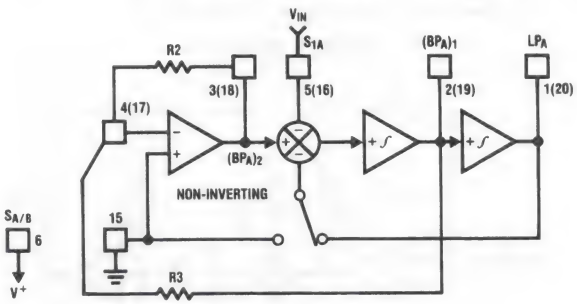

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FIGURE 8. MODE 1a

2.0 Modes of Operation (Continued)

MODE 1b: Notch 1, Bandpass, Lowpass Outputs: $f_{notch} = f_0$ (See *Figure 9*)

f_0 = center frequency of the complex pole pair

$$= \frac{f_{CLK}}{100} \times \sqrt{2} \text{ or } \frac{f_{CLK}}{50} \times \sqrt{2}$$

f_{notch} = center frequency of the imaginary zero pair = f_0.

H_{OLP} = Lowpass gain (as $f \rightarrow 0$) $= -\frac{R2}{2R1}$

H_{OBP} = Bandpass gain (at $f = f_0$) $= -\frac{R3}{R1}$

H_{ON} = Notch output gain as $\left.\begin{matrix} f \rightarrow 0 \\ f \rightarrow f_{CLK}/2 \end{matrix}\right\} = \frac{-R_2}{R_1}$

$$Q = \frac{f_0}{BW} = \frac{R3}{R2} \times \sqrt{2}$$

= quality factor of the complex pole pair

BW = the −3 dB bandwidth of the bandpass output.

Circuit dynamics:

$$H_{OLP} = \frac{H_{OBP}}{\sqrt{2}\,Q} \text{ or } H_{OBP} = H_{OLP} \times Q \times \sqrt{2}$$

$$H_{OBP} = \frac{H_{ON} \times Q}{\sqrt{2}}$$

$H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

MODE 2: Notch 2, Bandpass, Lowpass: $f_{notch} < f_0$ (See *Figure 10*)

f_0 = center frequency

$$= \frac{f_{CLK}}{100}\sqrt{\frac{R2}{R4} + 1} \text{ or } \frac{f_{CLK}}{50}\sqrt{\frac{R2}{R4} + 1}$$

$$f_{notch} = \frac{f_{CLK}}{100} \text{ or } \frac{f_{CLK}}{50}$$

Q = quality factor of the complex pole pair

$$= \frac{\sqrt{R2/R4 + 1}}{R2/R3}$$

H_{OLP} = Lowpass output gain (as $f \rightarrow 0$)

$$= -\frac{R2/R1}{R2/R4 + 1}$$

H_{OBP} = Bandpass output gain (at $f = f_0$) $= -R3/R1$

H_{ON_1} = Notch output gain (as $f \rightarrow 0$)

$$= -\frac{R2/R1}{R2/R4 + 1}$$

H_{ON_2} = Notch output gain $\left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right) = -R2/R1$

Filter dynamics: $H_{OBP} = Q\sqrt{H_{OLP}\,H_{ON_2}} = \sqrt{H_{ON_1}\,H_{ON_2}}$

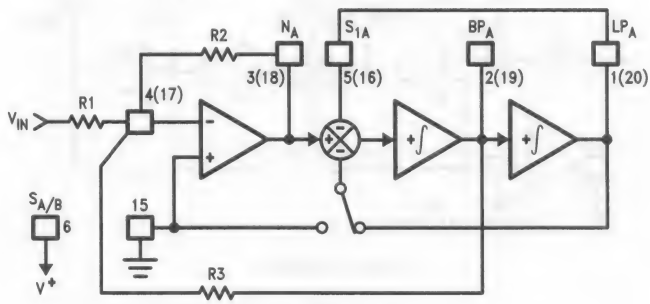

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FIGURE 9. MODE 1b

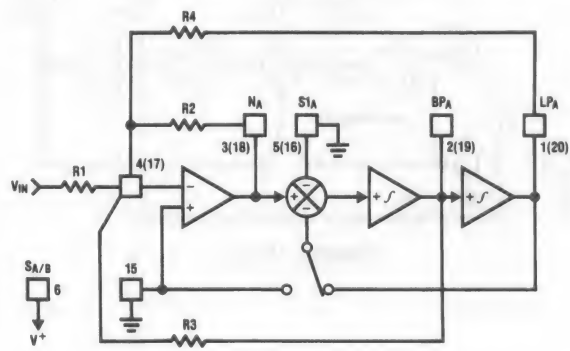

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FIGURE 10. MODE 2

2.0 Modes of Operation (Continued)

MODE 3: Highpass, Bandpass, Lowpass Outputs

(See *Figure 11*)

$$f_0 = \frac{f_{CLK}}{100} \times \sqrt{\frac{R2}{R4}} \text{ or } \frac{f_{CLK}}{50} \times \sqrt{\frac{R2}{R4}}$$

$$Q = \text{quality factor of the complex pole pair} = \sqrt{\frac{R2}{R4}} \times \frac{R3}{R2}$$

$$H_{OHP} = \text{Highpass gain} \left(\text{at } f \rightarrow \frac{f_{CLK}}{2} \right) = -\frac{R2}{R1}$$

$$H_{OBP} = \text{Bandpass gain (at } f = f_0) = -\frac{R3}{R1}$$

$$H_{OLP} = \text{Lowpass gain (as } f \rightarrow 0) = -\frac{R4}{R1}$$

Circuit dynamics: $\frac{R2}{R4} = \frac{H_{OHP}}{H_{OLP}}$; $H_{OBP} = \sqrt{H_{OHP} \times H_{OLP}} \times Q$

$H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

$H_{OHP(peak)} \cong Q \times H_{OHP}$ (for high Q's)

MODE 3a: HP, BP, LP and Notch with External Op Amp

(See *Figure 12*)

$$f_0 = \frac{f_{CLK}}{100} \times \sqrt{\frac{R2}{R4}} \text{ or } \frac{f_{CLK}}{50} \times \sqrt{\frac{R2}{R4}}$$

$$Q = \sqrt{\frac{R2}{R4}} \times \frac{R3}{R2}$$

$$H_{OHP} = -\frac{R2}{R1}$$

$$H_{OBP} = -\frac{R3}{R1}$$

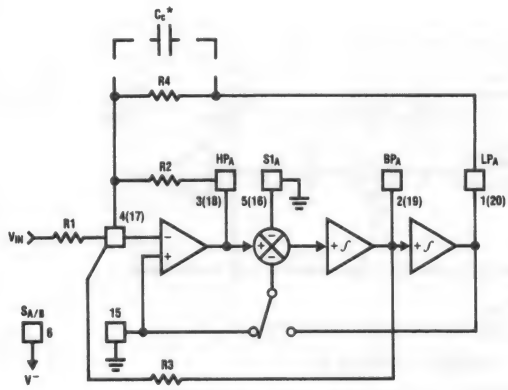
$$H_{OLP} = -\frac{R4}{R1}$$

$$f_n = \text{notch frequency} = \frac{f_{CLK}}{100} \sqrt{\frac{R_h}{R_l}} \text{ or } \frac{f_{CLK}}{50} \sqrt{\frac{R_h}{R_l}}$$

$$H_{ON} = \text{gain of notch at } f = f_0 = \left\| Q\left(\frac{R_g}{R_l} H_{OLP} - \frac{R_g}{R_h} H_{OHP}\right) \right\|$$

$$H_{n1} = \text{gain of notch (as } f \rightarrow 0) = \frac{R_g}{R_l} \times H_{OLP}$$

$$H_{n2} = \text{gain of notch} \left(\text{as } f \rightarrow \frac{f_{CLK}}{2} \right) = -\frac{R_g}{R_h} \times H_{OHP}$$



*In Mode 3, the feedback loop is closed around the input summing amplifier; the finite GBW product of this op amp causes a slight Q enhancement. If this is a problem, connect a small capacitor (10 pF − 100 pF) across R4 to provide some phase lead.

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FIGURE 11. MODE 3

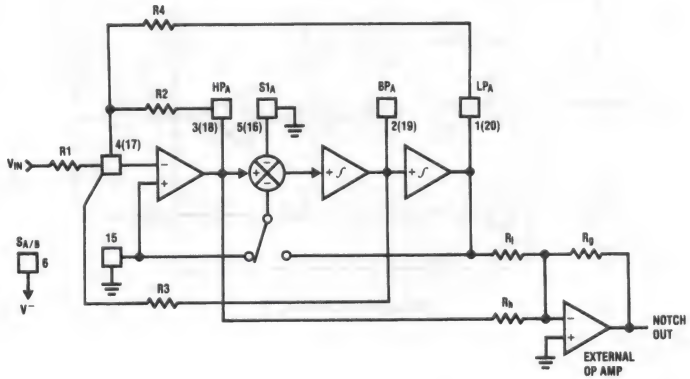

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FIGURE 12. MODE 3a

2.0 Modes of Operation (Continued)

MODE 4: Allpass, Bandpass, Lowpass Outputs

(See *Figure 13*)

f_0 = center frequency

$= \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$;

f_z* = center frequency of the complex zero $\approx f_0$

$Q = \frac{f_0}{BW} = \frac{R3}{R2}$;

Q_Z = quality factor of complex zero pair $= \frac{R3}{R1}$

For AP output make R1 = R2

H_{OAP}* = Allpass gain $\left(\text{at } 0 < f < \frac{f_{CLK}}{2}\right) = -\frac{R2}{R1} = -1$

H_{OLP} = Lowpass gain (as $f \rightarrow 0$)

$= -\left(\frac{R2}{R1} + 1\right) = -2$

H_{OBP} = Bandpass gain (at $f = f_0$)

$= -\frac{R3}{R2}\left(1 + \frac{R2}{R1}\right) = -2\left(\frac{R3}{R2}\right)$

Circuit dynamics: $H_{OBP} = (H_{OLP}) \times Q = (H_{OAP} + 1)Q$

*Due to the sampled data nature of the filter, a slight mismatch of f_z and f_0 occurs causing a 0.4 dB peaking around f_0 of the allpass filter amplitude response (which theoretically should be a straight line). If this is unacceptable, Mode 5 is recommended.

MODE 5: Numerator Complex Zeros, BP, LP

(See *Figure 14*)

$f_0 = \sqrt{1 + \frac{R2}{R4}} \times \frac{f_{CLK}}{100}$ or $\sqrt{1 + \frac{R2}{R4}} \times \frac{f_{CLK}}{50}$

$f_z = \sqrt{1 - \frac{R1}{R4}} \times \frac{f_{CLK}}{100}$ or $\sqrt{1 - \frac{R1}{R4}} \times \frac{f_{CLK}}{50}$

$Q = \sqrt{1 + R2/R4} \times \frac{R3}{R2}$

$Q_Z = \sqrt{1 - R1/R4} \times \frac{R3}{R1}$

H_{0z1} = gain at C.Z. output (as $f \rightarrow 0$ Hz) $\frac{-R2(R4 - R1)}{R1(R2 + R4)}$

H_{0z2} = gain at C.Z. output $\left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right) = \frac{-R2}{R1}$

$H_{OBP} = -\left(\frac{R2}{R1} + 1\right) \times \frac{R3}{R2}$

$H_{OLP} = -\left(\frac{R2 + R1}{R2 + R4}\right) \times \frac{R4}{R1}$

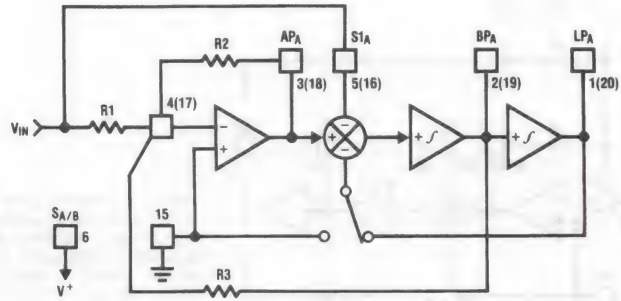

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FIGURE 13. MODE 4

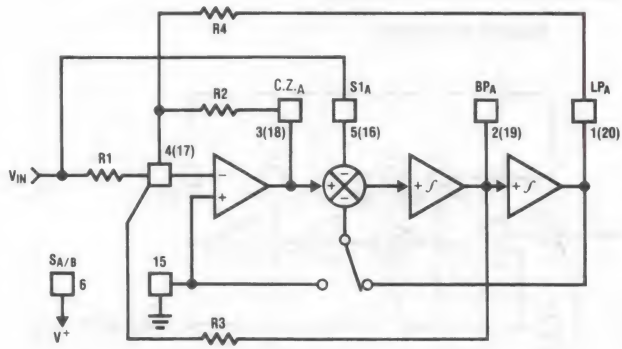

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FIGURE 14. MODE 5

2.0 Modes of Operation (Continued)

MODE 6a: Single Pole, HP, LP Filter (See *Figure 15*)

f_c = cutoff frequency of LP or HP output

$$= \frac{R2}{R3}\frac{f_{CLK}}{100} \text{ or } \frac{R2}{R3}\frac{f_{CLK}}{50}$$

$$H_{OLP} = -\frac{R3}{R1}$$

$$H_{OHP} = -\frac{R2}{R1}$$

MODE 6b: Single Pole LP Filter (Inverting and Non-Inverting) (See *Figure 16*)

f_c = cutoff frequency of LP outputs

$$\cong \frac{R2}{R3}\frac{f_{CLK}}{100} \text{ or } \frac{R2}{R3}\frac{f_{CLK}}{50}$$

$$H_{OLP_1} = 1 \text{ (non-inverting)}$$

$$H_{OLP_2} = -\frac{R3}{R2}$$

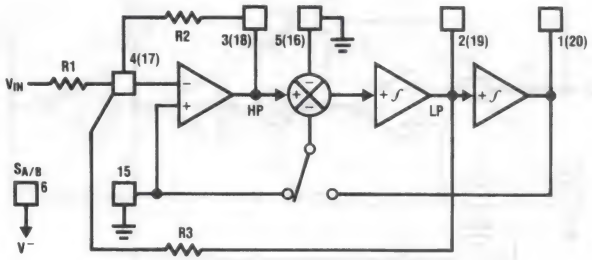

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FIGURE 15. MODE 6a

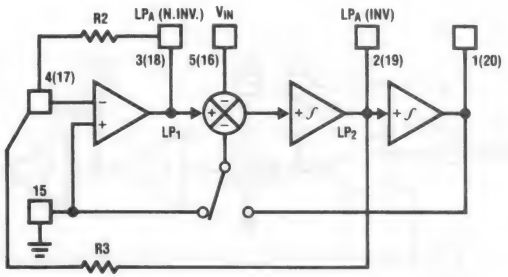

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FIGURE 16. MODE 6b

2.0 Modes of Operation (Continued)

MODE 6c: Single Pole, AP, LP Filter (See *Figure 17*)

$f_c = \frac{f_{CLK}}{50}$ or $\frac{f_{CLK}}{100}$

$H_{OAP} = 1$ (as $f \rightarrow 0$)

$H_{OAP} = -1$ (as $f \rightarrow f_{CLK}/2$)

$H_{OLP} = -2$

$R_1 = R_2 = R_3$

MODE 7: Summing Integrator (See *Figure 18*)

τ = integrator time constant

$= \frac{16}{f_{CLK}}$ or $\frac{8}{f_{CLK}}$

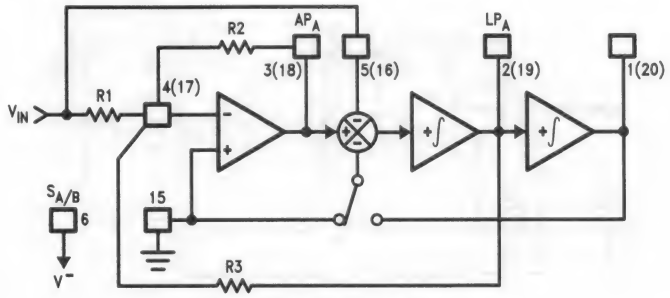

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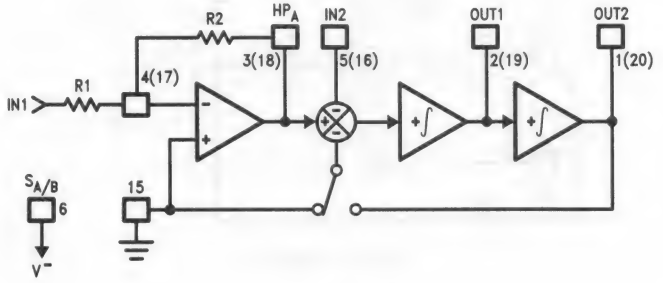
FIGURE 17. MODE 6c



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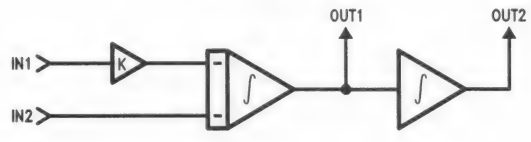
Equivalent Circuit



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$K = \frac{R_2}{R_1}$

$OUT1 = -\frac{k}{\tau}\int IN1\ dt - \frac{1}{\tau}\int IN2\ dt$

$OUT2 = \frac{1}{\tau}\int OUT1\ dt$

FIGURE 18. MODE 7

2.0 Modes of Operation (Continued)

TABLE I. Summary of Modes. Realizable filter types (e.g. low-pass) denoted by asterisks. Unless otherwise noted, gains of various filter outputs are inverting and adjustable by resistor ratios.

Mode	BP	LP	HP	N	AP	Number of Resistors	Adjustable f_{CLK}/f_0	Notes
1	*	*		*		3	No	
1a	(2) $H_{OBP1} = -Q$ $H_{OBP2} = +1$	$H_{OLP} = +1$				2	No	May need input buffer. Poor dynamics for high Q.
1b	*	*		*		3	No	Useful for high frequency applications.
2	*	*		*		3	Yes (above $f_{CLK}/50$ or $f_{CLK}/100$)	
3	*	*	*			4	Yes	Universal State-Variable Filter. Best general-purpose mode.
3a	*	*	*	*		7	Yes	As above, but also includes resistor-tuneable notch.
4	*	*			*	3	No	Gives Allpass response with $H_{OAP} = -1$ and $H_{OLP} = -2$.
5	*	*			*	4	Yes	Gives flatter allpass response than above if $R_1 = R_2 = 0.02R_4$.
6a		*	*			3	Yes	Single pole.
6b		(2) $H_{OLP1} = +1$ $H_{OLP2} = \frac{-R3}{R2}$				2	Yes	Single pole.
6c		*			*	3	No	Single pole.
7						2	Yes	Summing integrator with adjustable time constant.

3.0 Applications Information

The LMF100 is a general purpose dual second-order state variable filter whose center frequency is proportional to the frequency of the square wave applied to the clock input (f_{CLK}). The various clocking options are summarized in the following table.

Clocking Options

Power Supply	Clock Levels	LSh	V_D^+
−5V and +5V	TTL (0V to +5V)	0V	+5V
−5V and +5V	CMOS (−5V to +5V)	0V	+5V
0V and 10V	TTL (0V to 5V)	0V	+10V
0V and 10V	CMOS (0V to +10V)	+5V	+10V
−2.5V and +2.5V	CMOS (−2.5V to +2.5V)	0V	+2.5V
0V and 5V	TTL (0V to +5V)	0V	0V
0V and 5V	CMOS (0V to +5V)	+2.5V	+5V

By connecting pin 12 to the appropriate dc voltage, the filter center frequency, f_0, can be made equal to either $f_{CLK}/100$ or $f_{CLK}/50$. f_0 can be very accurately set (within ±0.6%) by using a crystal clock oscillator, or can be easily varied over a wide frequency range by adjusting the clock frequency. If desired, the f_{CLK}/f_0 ratio can be altered by external resistors as in *Figures 10, 11, 12, 13, 14, 15* and *16*. This is useful when high-order filters (greater than two) are to be realized by cascading the second-order sections. This allows each stage to be stagger tuned while using only one clock. The filter Q and gain are set by external resistor ratios.

All of the five second-order filter types can be built using either section of the LMF100. These are illustrated in *Figures 1* through *5* along with their transfer functions and some related equations. *Figure 6* shows the effect of Q on the shapes of these curves.

3.0 Applications Information (Continued)

3.1 DESIGN EXAMPLE

In order to design a filter using the LMF100, we must define the necessary values of three parameters for each second-order section: f_0, the filter section's center frequency; H_0, the passband gain; and the filter's Q. These are determined by the characteristics required of the filter being designed.

As an example, let's assume that a system requires a fourth-order Chebyshev low-pass filter with 1 dB ripple, unity gain at dc, and 1000 Hz cutoff frequency. As the system order is four, it is realizable using both second-order sections of an LMF100. Many filter design texts (and National's Switched Capacitor Filter Handbook) include tables that list the characteristics (f_0 and Q) of each of the second-order filter sections needed to synthesize a given higher-order filter. For the Chebyshev filter defined above, such a table yields the following characteristics:

$f_{0A} = 529$ Hz $\quad Q_A = 0.785$

$f_{0B} = 993$ Hz $\quad Q_B = 3.559$

For unity gain at dc, we also specify:

$H_{0A} = 1$

$H_{0B} = 1$

The desired clock-to-cutoff-frequency ratio for the overall filter of this example is 100 and a 100 kHz clock signal is available. Note that the required center frequencies for the two second-order sections will not be obtainable with clock-to-center-frequency ratios of 50 or 100. It will be necessary to adjust $\frac{f_{CLK}}{f_0}$ externally. From Table I, we see that Mode 3 can be used to produce a low-pass filter with resistor-adjustable center frequency.

In most filter designs involving multiple second-order stages, it is best to place the stages with lower Q values ahead of stages with higher Q, especially when the higher Q is greater than 0.707. This is due to the higher relative gain at the center frequency of a higher-Q stage. Placing a stage with lower Q ahead of a higher-Q stage will provide some attenuation at the center frequency and thus help avoid clipping of signals near this frequency. For this example, stage A has the lower Q (0.785) so it will be placed ahead of the other stage.

For the first section, we begin the design by choosing a convenient value for the input resistance: $R_{1A} = 20$k. The absolute value of the passband gain H_{OLPA} is made equal to 1 by choosing R_{4A} such that: $R_{4A} = -H_{OLPA}R_{1A} = R_{1A} = 20$k. If the 50/100/CL pin is connected to mid-supply for nominal 100:1 clock-to-center-frequency ratio, we find R_{2A} by:

$$R_{2A} = R_{4A}\frac{f_{0A}^2}{(f_{CLK}/100)^2} = 2 \times 10^4 \times \frac{(529)^2}{(1000)^2} = 5.6\text{k and}$$

$$R_{3A} = Q_A\sqrt{R_{2A}R_{4A}} = 0.785\sqrt{5.6 \times 10^3 \times 2 \times 10^4} = 8.3\text{k}$$

The resistors for the second section are found in a similar fashion:

$R_{1B} = 20$k

$R_{4B} = R_{1B} = 20$k

$$R_{2B} = R_{4B}\frac{f_{0B}^2}{(f_{CLK}/100)^2} = 20\text{k}\frac{(993)^2}{(1000)^2} = 19.7\text{k}$$

$$R_{3B} = Q_B\sqrt{R_{2B}R_{4B}} = 3.559\sqrt{1.97 \times 10^4 \times 2 \times 10^4} = 70.6\text{k}$$

The complete circuit is shown in *Figure 19* for split ±5V power supplies. Supply bypass capacitors are highly recommended.

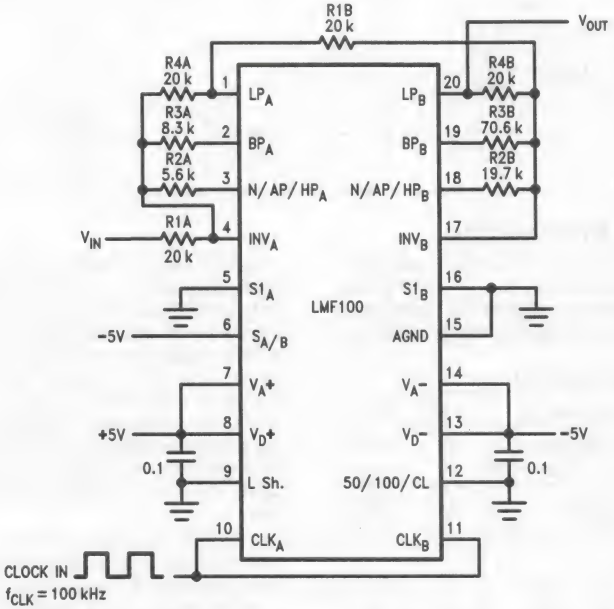

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FIGURE 19. Fourth-order Chebyshev low-pass filter from example in 3.1. ±5V power supply. 0V–5V TTL or ±5V CMOS logic levels.

3.0 Applications Information (Continued)

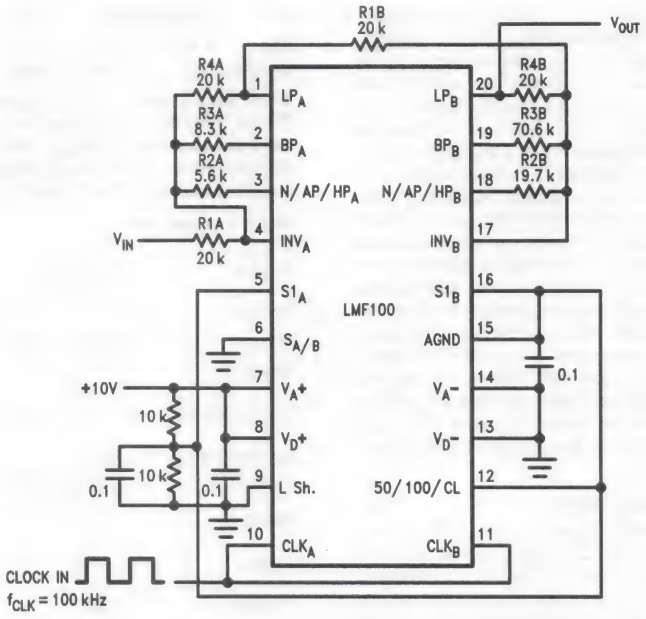
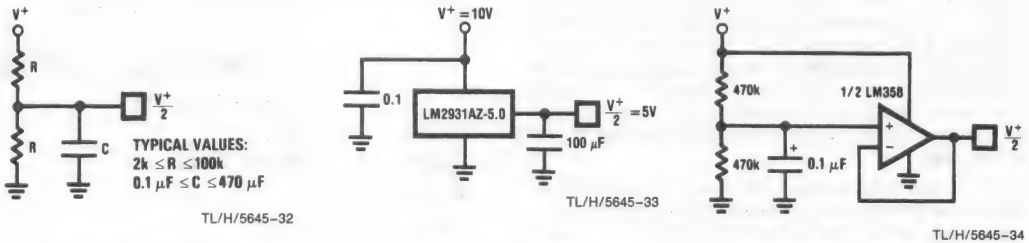


FIGURE 20. Fourth-order Chebyshev low-pass filter from example in 3.1. Single + 10V power supply. 0V–5V TTL logic levels. Input signals should be referred to half-supply or applied through a coupling capacitor.



(a) Resistive Divider with Decoupling Capacitor

(b) Voltage Regulator

(c) Operational Amplifier with Divider

FIGURE 21. Three Ways of Generating $\frac{V^+}{2}$ for Single-Supply Operation

3.0 Applications Information (Continued)

3.2 SINGLE SUPPLY OPERATION

The LMF100 can also operate with a single-ended power supply. *Figure 20* shows the example filter with a single-ended power supply. V_A^+ and V_D^+ are again connected to the positive power supply (4 to 15 volts), and V_A^- and V_D^- are connected to ground. The A_{GND} pin must be tied to $V^+/2$ for single supply operation. This half-supply point should be very "clean", as any noise appearing on it will be treated as an input to the filter. It can be derived from the supply voltage with a pair of resistors and a bypass capacitor *(Figure 21a)*, or a low-impedance half-supply voltage can be made using a three-terminal voltage regulator or an operational amplifier *(Figures 21b* and *21c)*. The passive resistor divider with a bypass capacitor is sufficient for many applications, provided that the time constant is long enough to reject any power supply noise. It is also important that the half-supply reference present a low impedance to the clock frequency, so at very low clock frequencies the regulator or op-amp approaches may be preferable because they will require smaller capacitors to filter the clock frequency. The main power supply voltage should be clean (preferably regulated) and bypassed with 0.1 μF.

3.3 DYNAMIC CONSIDERATIONS

The maximum signal handling capability of the LMF100, like that of any active filter, is limited by the power supply voltages used. The amplifiers in the LMF100 are able to swing to within about 1 volt of the supplies, so the input signals must be kept small enough that none of the outputs will exceed these limits. If the LMF100 is operating on ± 5 volts, for example, the outputs will clip at about $8V_{p-p}$. The maximum input voltage multiplied by the filter gain should therefore be less than $8V_{p-p}$.

Note that if the filter Q is high, the gain at the lowpass or highpass outputs will be much greater than the nominal filter gain *(Figure 6)*. As an example, a lowpass filter with a Q of 10 will have a 20 dB peak in its amplitude response at f_0. If the nominal gain of the filter (H_{OLP}) is equal to 1, the gain at f_0 will be 10. The maximum input signal at f_0 must therefore be less than 800 mV_{p-p} when the circuit is operated on ± 5 volt supplies.

Also note that one output can have a reasonable small voltage on it while another is saturated. This is most likely for a circuit such as the notch in Mode 1 *(Figure 7)*. The notch output will be very small at f_0, so it might appear safe to apply a large signal to the input. However, the bandpass will have its maximum gain at f_0 and can clip if overdriven. If one output clips, the performance at the other outputs will be degraded, so avoid overdriving any filter section, even ones whose outputs are not being directly used. Accompanying *Figures 7* through *17* are equations labeled "circuit dynamics", which relate the Q and the gains at the various outputs. These should be consulted to determine peak circuit gains and maximum allowable signals for a given application.

3.4 OFFSET VOLTAGE

The LMF100's switched capacitor integrators have a slightly higher input offset voltage than found in a typical continuous time active filter integrator. Because of National's new LMCMOS process and new design techniques the internal offsets have been minimized, compared to the industry standard MF10. *Figure 22* shows an equivalent circuit of the LMF100 from which the output dc offsets can be calculated.

Typical values for these offsets with $S_{A/B}$ tied to V^+ are:

V_{OS1} = opamp offset = ± 5 mV

V_{OS2} = ± 30 mV at 50:1 or 100:1

V_{OS3} = ± 15 mV at 50:1 or 100:1

When $S_{A/B}$ is tied to V^-, V_{OS2} will approximately halve. The dc offset at the BP output is equal to the input offset of the lowpass integrator (V_{OS3}). The offsets at the other outputs depend on the mode of operation and the resistor ratios, as described in the following expressions.

Mode 1 and Mode 4

$$V_{OS(N)} = V_{OS1}\left(\frac{1}{Q} + 1 + \left\|H_{OLP}\right\|\right) - \frac{V_{OS3}}{Q}$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N)} - V_{OS2}$$

Mode 1a

$$V_{OS(N.INV.BP)} = \left(1 + \frac{1}{Q}\right)V_{OS1} - \frac{V_{OS3}}{Q}$$

$$V_{OS(INV.BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N.INV.BP)} - V_{OS2}$$

Mode 1b

$$V_{OS(N)} = V_{OS1}\left(1 + \frac{R2}{R3} + \frac{R2}{R1}\right) - \frac{R2}{R3}V_{OS3}$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = \frac{V_{OS(N)}}{2} - \frac{V_{OS2}}{2}$$

Mode 2 and Mode 5

$$V_{OS(N)} = \left(\frac{R2}{Rp} + 1\right)V_{OS1} \times \frac{1}{1 + R2/R4} + V_{OS2}\frac{1}{1 + R4/R2} - \frac{V_{OS3}}{Q\sqrt{1 + R2/R4}}$$

$$R_p = R1 \| R3 \| R4$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N)} - V_{OS2}$$

Mode 3

$$V_{OS(HP)} = V_{OS2}$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS1}\left[1 + \frac{R4}{R_p}\right] - V_{OS2}\left(\frac{R4}{R2}\right) - V_{OS3}\left(\frac{R4}{R3}\right)$$

$$R_p = R1 \| R2 \| R3$$

Mode 6a and 6c

$$V_{OS(HP)} = V_{OS2}$$

$$V_{OS(LP)} = V_{OS1}\left(1 + \frac{R_3}{R_2} + \frac{R_3}{R_1}\right) - \frac{R_3}{R_2}V_{OS2}$$

Mode 6b

$$V_{OS(LP\ (N.INV))} = V_{OS2}$$

$$V_{OS(LP\ (INV))} = V_{OS1}\left(1 + \frac{R_3}{R_2}\right) - \frac{R_3}{R_2}V_{OS2}$$

3.0 Applications Information (Continued)

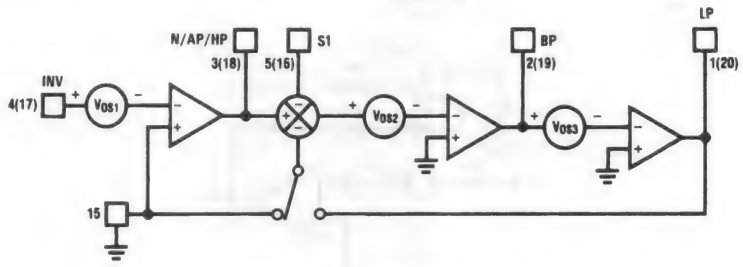


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FIGURE 22. Offset Voltage Sources

In many applications, the outputs are ac coupled and dc offsets are not bothersome unless large signals are applied to the filter input. However, larger offset voltages will cause clipping to occur at lower ac signal levels, and clipping at any of the outputs will cause gain nonlinearities and will change f_0 and Q. When operating in Mode 3, offsets can become excessively large if R_2 and R_4 are used to make f_{CLK}/f_0 significantly higher than the nominal value, especially if Q is also high.

For example, *Figure 23* shows a second-order 60 Hz notch filter. This circuit yields a notch with about 40 dB of attenuation at 60 Hz. A notch is formed by subtracting the bandpass output of a mode 3 configuration from the input using the unused side B opamp. The Q is 10 and the gain is 1 V/V in the passband. However, $f_{CLK}/f_0 = 1000$ to allow for a wide input spectrum. This means that for pin 12 tied to ground (100:1 mode), R4/R2 = 100. The offset voltage at the lowpass output (LP) will be about 3V. However, this is an extreme case and the resistor ratio is usually much smaller. Where necessary, the offset voltage can be adjusted by using the circuit of *Figure 24*. This allows adjustment of V_{OS1}, which will have varying effects on the different outputs as described in the above equations. Some outputs cannot be adjusted this way in some modes, however ($V_{OS(BP)}$ in modes 1a and 3, for example).

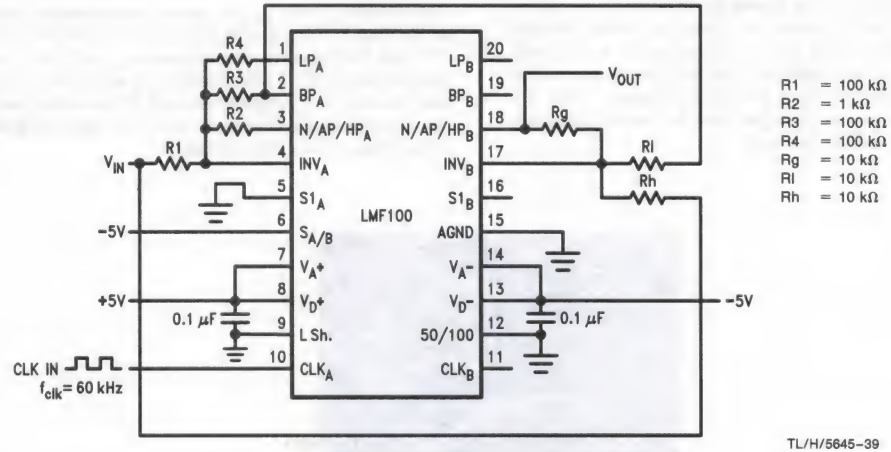

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FIGURE 23. Second-Order Notch Filter

3.0 Applications Information (Continued)

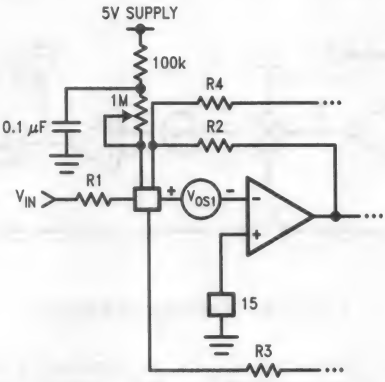

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FIGURE 24. Method for Trimming V_{OS}

3.5 SAMPLED DATA SYSTEM CONSIDERATIONS

The LMF100 is a sampled data filter, and as such, differs in many ways from conventional continuous-time filters. An important characteristic of sampled-data systems is their effect on signals at frequencies greater than one-half the sampling frequency. (The LMF100's sampling frequency is the same as its clock frequency.) If a signal with a frequency greater than one-half the sampling frequency is applied to the input of a sampled data system, it will be "reflected" to a frequency less than one-half the sampling frequency. Thus, an input signal whose frequency is $f_s/2 + 100$ Hz will cause the system to respond as though the input frequency was $f_s/2 - 100$ Hz. This phenomenon is known as "aliasing", and can be reduced or eliminated by limiting the input signal spectrum to less than $f_s/2$. This may in some cases require the use of a bandwidth-limiting filter ahead of the LMF100 to limit the input spectrum. However, since the clock frequency is much higher than the center frequency, this will often not be necessary.

Another characteristic of sampled-data circuits is that the output signal changes amplitude once every sampling period, resulting in "steps" in the output voltage which occur at the clock rate *(Figure 25)*. If necessary, these can be "smoothed" with a simple R-C low-pass filter at the LMF100 output.

The ratio of f_{CLK} to f_c (normally either 50:1 or 100:1) will also affect performance. A ratio of 100:1 will reduce any aliasing problems and is usually recommended for wideband input signals. In noise-sensitive applications, a ratio of 100:1 will result in 3 dB lower output noise for the same filter configuration.

The accuracy of the f_{CLK}/f_0 ratio is dependent on the value of Q. This is illustrated in the curves under the heading "Typical Performance Characteristics". As Q is changed, the true value of the ratio changes as well. Unless the Q is low, the error in f_{CLK}/f_0 will be small. If the error is too large for a specific application, use a mode that allows adjustment of the ratio with external resistors.

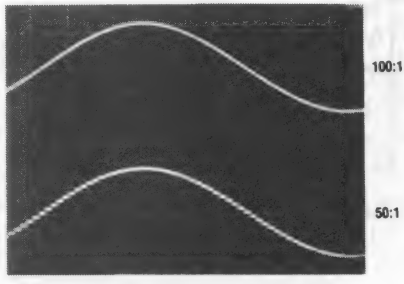

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FIGURE 25. The Sampled-Data Output Waveform

National Semiconductor

LMF120 Mask-Programmable Switched-Capacitor Active Filter System

General Description

The LMF120 is a mask-programmable switched-capacitor filter capable of realizing virtually any filter response up to twelve poles using six independent biquad blocks. It is customized to meet specific application requirements through the use of automated design techniques. Circuit realization occurs during the final metal-mask stage of the manufacturing process.

Three sample-and-hold inputs and three buffered outputs allow one, two, or three independent filters on a single chip. Each of the filters may be any type: high-pass, low-pass, all-pass, bandpass, or notch.

The center or cutoff frequency of each filter is determined by the clock frequency. The clock signal can be supplied by an external source, or it can be generated by the internal oscillator, using an external crystal and two capacitors. An on-board programmable divider chain can divide the clock input frequency by up to 256 so that each on-chip filter can have a different cutoff/center frequency. Accuracy is enhanced by close matching of the internal components: the ratio of the clock frequency to the center/corner frequency is typically accurate to ±0.5%, and is guaranteed to ±1.5% over the full temperature range.

The customization process is initiated by submitting transfer functions, pole and zero locations, or band diagrams to National Semiconductor. A worksheet is included in this datasheet, which can be returned to National Semiconductor for an initial evaluation. Each filter is computer-optimized to best meet the requested specifications, and computer simulations are produced for approval before prototyping begins.

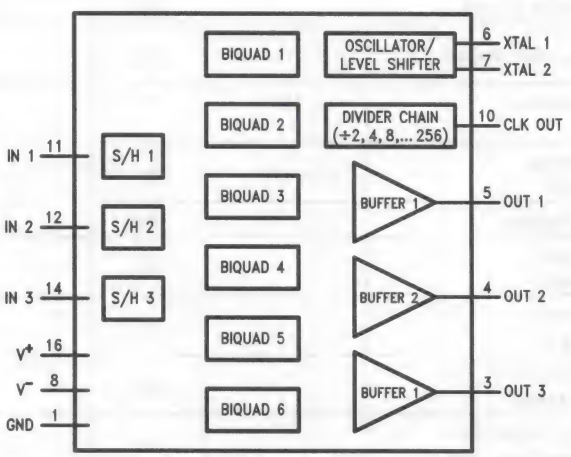
Features

- Mask-programmable for virtually any filter response
- All filter types (low-pass, high-pass, bandpass, notch, all-pass)
- All filter approximations (Butterworth, Chebyshev, Elliptic, Bessel, etc.)
- Up to 12 poles and right-half-plane zeros in one 16-pin package
- One, two, or three filters per package
- Wide Q range: up to 100 per biquad
- Choice of internal or external clock
- No external components other than clock or crystal and two capacitors
- Programmable clock divider: ÷2 to ÷256
- Center frequency accuracy: ±1.5% over temperature
- Supply voltage range: ±2V to ±7.5V or +4V to +14V

Applications

- Anti-alias filters
- Real-time audio analyzers
- Biomedical instrumentation
- Cellular telephones

Simplified Block Diagram



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Absolute Maximum Ratings (Notes 1 & 3)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Total Supply Voltage ($V^+ - V^-$)	−0.3V to +16V
Voltage at Any Pin	V^- − 0.3V to V^+ + 0.3V
Input Current per Pin (Note 10)	±5 mA
Total Input Current (Note 10)	±20 mA
Lead Temp. (Soldering 10 sec.)	
Dual-In-Line Package (Plastic)	300°C
Surface Mount Pkg. (Note 4)	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C
Power Dissipation (Note 5)	500 mW
Maximum Junction Temperature	150°C
Storage Temperature Range	−65°C to +150°C
ESD Susceptibility (Note 6)	2000V

Operating Ratings (Notes 2 & 3)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
LMF120CCN, LMF120CCV	$0°C \le T_A \le +70°C$
LMF120CIJ	$-40°C \le T_A \le +85°C$
LMF120CMJ	$-55°C \le T_A \le +125°C$
Supply Voltage ($V^+ - V^-$)	4.0V to 14V

Filter Electrical Characteristics

Because the LMF120's performance characteristics vary depending on the programming mask configuration, many of the specifications listed in this section are given only as typical values. These are intended to serve as guidelines for assessing the capabilities of the IC and the feasibility of a desired filter response. Specific filter performance data (obtained by computer simulation) will be supplied by National Semiconductor after the desired characteristics for the particular filter implementation have been defined. Test frequencies and attenuation values appropriate to the application can then be chosen. The following specifications apply for $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for T_{Min} to T_{Max}**; all other limits apply for $T_A = T_J = 25°C$.

Symbol	Parameter	Conditions	Typical (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Units (Limit)
f_{CLK}	Filter Clock Frequency		10 1.5			Hz (Min) MHz (Max)
f_{CLKIN}	Clock Input Frequency (Logic Circuitry Only)	Pin 6 or 7	4			MHz (Max)
f_0	Center or Cutoff Frequency		0.1 100			Hz (Min) kHz (Max)
f_{CLK}/f_0	Filter Clock-to-Center-Frequency Ratio Range (Each Biquad)		10 500			Hz/Hz (Min) Hz/Hz (Max)
$\Delta f_{CLK}/f_0$	Filter Clock-to-Center-Frequency Accuracy (Each Biquad)		±0.5			% (Max)
H_0	Passband Gain Error		±0.2			dB (Max)
Q	Filter "Q" (Each Biquad)		100			(Max)
$\frac{\Delta Q}{Q}$	Q Accuracy (Each Biquad)	$0.5 \le Q \le 30$	±2			%
$f_0 \times Q$	Center Frequency-Q Product	$Q \le 100$	1			MHz
	Dynamic Range (Each Biquad)	(Note 11)	80			dB
	Clock Feedthrough		10			mVrms
V_{OS}	Offset Voltage (Each Biquad)		70			mV
I_{SBQ}	Supply Current (Each Biquad)		0.4			mA
I_{SSH}	Supply Current (Each Input Sample-and-Hold)		0.3			mA
I_S	Total Supply Current (All Circuit Blocks Connected)		10			mA

Output Buffer Electrical Characteristics

The following specifications apply for $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for T_{Min} to T_{Max}**; all other limits apply for $T_A = T_J = 25°C$.

Symbol	Parameter	Conditions	Typical (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Units (Limit)
V_O	Output Voltage Swing	$R_L = 5\ k\Omega$		$V^+ - 1.0$ $V^- + 1.0$		V (Min) V (Max)
SR	Slew Rate		1.0			V/μs
C_L	Maximum Capacitive Load		200			pF
GBW	Gain-Bandwidth Product		1.0			MHz
I_S	Supply Current per Buffer		0.8			mA

Logic Input and Output Electrical Characteristics

The following specifications apply for $V^+ = +5V$ and $V^- = -5V$ unless otherwise specified. **Boldface limits apply for T_{Min} to T_{Max}**; all other limits apply for $T_A = T_J = +25°C$.

Symbol	Parameter		Conditions	Typical (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Units (Limit)
V_{IH} V_{IL}	Pin 7 CMOS Clock Input Voltage (Notes 12 and 13)	Logical "1" Logical "0"	$V^+ = 5V, V^- = -5V$		+3.0 −3.0		V (Min) V (Max)
V_{IH} V_{IL}		Logical "1" Logical "0"	$V^+ = 10V, V^- = 0V$		+8.0 +2.0		V (Min) V (Max)
V_{IH} V_{IL}		Logical "1" Logical "0"	$V^+ = 2.5V, V^- = -2.5V$		+1.5 −1.5		V (Min) V (Max)
V_{IH} V_{IL}		Logical "1" Logical "0"	$V^+ = 5V, V^- = 0V$		+4.0 +1.0		V (Min) V (Max)
V_{IH} V_{IL}	Pin 6 TTL Clock Input Voltage (Notes 12 and 13)	Logical "1" Logical "0"	$V^+ = 5V, V^- = -5V$		+2.0 +0.8		V (Min) V (Max)
V_{IH} V_{IL}		Logical "1" Logical "0"	$V^+ = 10V, V^- = 0V$		+2.0 +0.8		V (Min) V (Max)
V_{IH} V_{IL}		Logical "1" Logical "0"	$V^+ = 5V, V^- = 0V$		+2.0 +0.8		V (Min) V (Max)
V_{OH}	Clock Output	Logical "1"	$I_{OUT} = -1\ mA$		$V^+ - 1.0$		V (Min)
V_{OL}	Clock Output	Logical "0"	$I_{OUT} = +1\ mA$		$V^- + 1.0$		V (Max)
I_{IN}	Input Current	XTAL1, XTAL2			±10		μA (Max)

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur.

Note 2: Operating Ratings indicate conditions for which the device is intended to be functional. These ratings do not guarantee specific performance limits, however. For guaranteed specifications and test conditions, see the Electrical Characteristics. The guaranteed specifications apply only for the test conditions listed. Some performance characteristics may degrade when the device is not operated under the listed test conditions.

Note 3: All voltages are measured with respect to GND unless otherwise specified.

Note 4: See AN450 "Surface Mounting Methods and Their Effect on Product Reliability" or the section titled "Surface Mount" found in any current Linear Data Book for other methods of soldering surface mount devices.

Note 5: The maximum power dissipation must be derated at elevated temperatures and is a function of T_{Jmax}, Θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{Jmax} - T_A)/\Theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For guaranteed operation, $T_{Jmax} = 125°C$. The typical thermal resistance (Θ_{JA}) of the LMF120N when board-mounted is 51°C/W. Θ_{JA} is typically 52°C/W for the LMF120J, and 86°C/W for the LMF120V.

Note 6: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 7: Typicals are at $T_J = 25°C$ and represent the most likely parametric norm.

Note 8: Tested Limits are guaranteed and 100% tested.

Note 9: Design Limits are guaranteed, but not 100% tested.

Note 10: When the input voltage (V_{IN}) at any pin exceeds the power supplies ($V_{IN} < V^-$ or $V_{IN} > V^+$), the current at that pin should be limited to 5 mA. The 20 mA maximum package input current rating limits the number of pins that can safely exceed the power supplies with an input current of 5 mA to four.

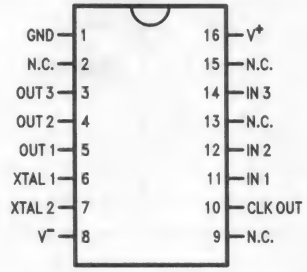
Note 11: Dynamic range is defined as the ratio of the tested minimum output voltage swing to the wideband noise over a 20 kHz bandwidth.

Note 12: Each custom version of the LMF120 will be tested at only one power supply voltage, which will be chosen to correspond to the application for which it is intended.

Note 13: Only one clock input will be active for any given version of the LMF120. Therefore, a device wil be tested for either TTL or CMOS clock input threshold, whichever is appropriate.

Connection Diagrams

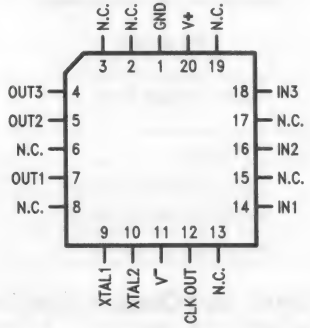
Dual-In-Line Package



TL/H/10353–2

Order Number LMF120CIJ, LMF120CMJ, LMF120CCN
See NS Package Number J16A or N16E

Plastic Chip Carrier Package

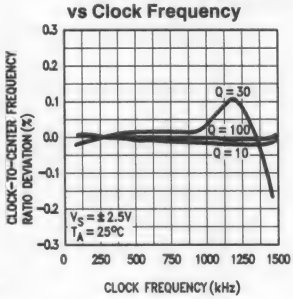


TL/H/10353–3

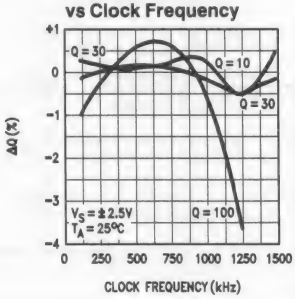
Order Number LMF120CCV
See NS Package Number V20A

Typical Performance Characteristics

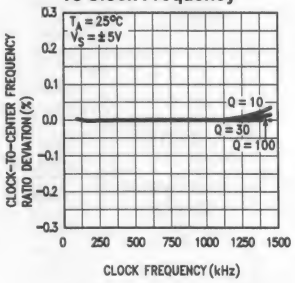
Biquad Clock-to-Center Frequency Ratio Deviation vs Clock Frequency



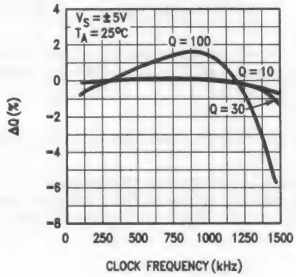
Biquad Q Deviation vs Clock Frequency



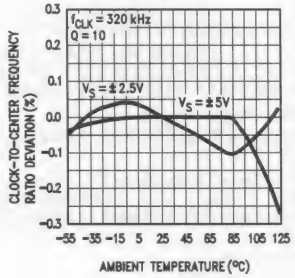
Biquad Clock-to-Center Frequency Ratio Deviation vs Clock Frequency



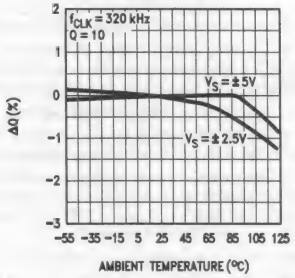
Biquad Q Deviation vs Clock Frequency



Biquad Clock-to-Center-Frequency Ratio Deviation vs Temperature



Biquad Q Deviation vs Temperature



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Pin Description

GND (Pin 1) This is the analog ground reference for the LMF120. In split supply applications, GND should be connected to the system ground. When operating the LMF120 from a single positive power supply voltage, pin 1 should be connected to a "clean" reference voltage midway between V^+ and V^-.

N.C. (Pins 2, 9, 13, & 15) These pins are not connected to the internal circuitry.

OUT3 (Pin 3), OUT2 (Pin 4), OUT1 (Pin 5) These are the outputs of the buffer amplifiers. Depending on the filter configuration, one, two, or all three of these outputs may be used.

XTAL1 (Pin 6) This is the crystal oscillator input pin. When using the internal oscillator, the crystal should be tied between XTAL1 and XTAL2. XTAL1 can also be used as an input for an external TTL-level clock.

XTAL2 Pin 7) This is the output of the internal crystal oscillator. When using the internal oscillator, the crystal should be tied between XTAL1 and XTAL2. XTAL2 can also be used as an input for an external CMOS logic-compatible clock swinging from V^+ to V^-.

V^- (Pin 8) This is the negative power supply pin. It should be bypassed with at least a 0.1 μF ceramic capacitor. For single-supply operation, connect this pin to system ground.

CLOCK OUT (Pin 10) This is the clock output pin. It can drive the clock inputs of additional filters or other components. The clock output signal swings from V^+ to V^-. This pin can be mask-programmed to supply an output at the same frequency as the internal oscillator or external clock input, or at any output frequency available from the internal divider chain.

INPUT1 (Pin 11), INPUT2 (Pin 12), INPUT3 (Pin 14) These are the inputs to the filter. When necessary (in notch filters, for example), the input pins are connected to the internal sample-hold circuits.

V^+ (Pin 16) This is the positive power supply pin. It should be bypassed with at least a 0.1 μF ceramic capacitor.

Functional Description

Each of the six internal biquad switched-capacitor filter sections (shown in detail in *Figure 1*) can have a characteristic equation of the form:

$$V_{OUT}(s) = \frac{S^2V_{HP} - V_{BP2}b_1s + V_{LP}b_0}{s^2 + a_1s + a_0} \quad (1)$$

or:

$$V_{OUT}(s) = \frac{-(s^2V_{HP} + V_{BP1}b_1s)}{s^2 + a_1s + a_0} \quad (2)$$

Note that by proper choice of coefficients and input connections, any type of filter response (low-pass, high-pass, bandpass, notch, or all-pass) can be obtained. For example, a notch filter can be realized by connecting the input signal to V_{HP} and V_{LP}. An all-pass filter can be realized by connecting the input signal to V_{HP}, V_{LP}, and V_{BP2}. Coefficients are controlled by the metal mask, which determines the values of the internal capacitors and the interconnections between the filter stages, sample-holds, and output buffers. By appropriate design of the metal mask, the biquad sections can be cascaded to form high-order filters.

The center or cutoff frequency is proportional to the filter clock frequency. The ratio of the clock frequency to the center frequency (f_{CLK}:f_0) is programmable with virtually infinite resolution over a range of 10:1 to 500:1, although clock-to-center-frequency ratios in the 50:1 to 100:1 range usually give the best performance.

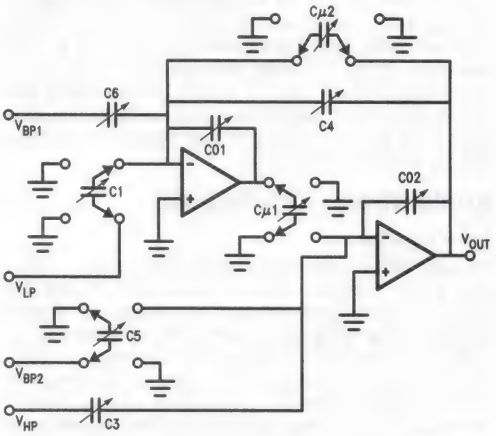

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FIGURE 1. Single Biquad Structure. There are six of these second-order blocks within the LMF120. Any of the biquad blocks can realize a low-pass, high-pass, bandpass, notch, or all-pass response.

The LMF120 contains three input sample-and-hold circuits. These are used only when necessary—in a notch filter, for example, where a sampled signal is summed with a continuous signal within the biquad. The result of such a summation would contain a residual signal equal to the difference between the sampled waveform and its continuous version. This residual would place a limit on the notch filter's effectiveness. The sample-and-hold ensures that the "continuous" signal path in the biquad (from V_{HP} to V_{OUT}) carries a sampled signal, thus improving the notch's performance.

In addition to three input pins, the LMF120 has three output buffer amplifiers, allowing one package to contain up to three independent filters. The total number of poles can be any number up to twelve, so, for example, a single LMF120 could perform the function of a 6th-order low-pass, a 2nd-order bandpass, and a 4th-order high-pass filter simultaneously.

Functional Description (Continued)

Clock Circuitry

The LMF120's clock input circuitry can be mask-programmed to accept an external TTL or CMOS-level clock, or to serve as a self-contained oscillator with the addition of an external crystal and two capacitors (see *Figure 6*). The clock signal can directly drive the biquad sections (if the frequency is appropriate), or its frequency can be divided by 2^n, where n is an integer between 1 and 8 ($\div 2$, $\div 4$, $\div 8$, ... $\div 256$). If necessary, each biquad section can obtain its clock signal from a different divider tap.

The Clock Output pin can be programmed to supply additional LMF120s or other circuits with a clock signal whose frequency is equal to either the clock input frequency, or the frequency at any of the divider taps.

Power Consumption

Because the LMF120 is a CMOS integrated circuit, its power consumption is low. To further reduce power consumption, any unused sample-and-holds and buffer amplifiers are shut down when fewer than three filters are required. (For example, a single 12th-order notch filter would need only one sample-and-hold and one buffer.) Unused biquad sections (if any) are shut down as well. For low-frequency applications, the internal current drain can be reduced by about 30% for further power savings.

Applications Information

Power Supplies

The LMF120 can operate from supply voltages ($V^+ - V^-$) ranging from 4.0V up to 14V, but the choice of supply voltage can affect circuit performance. The IC depends on MOS switches for its operation. All such switches have inherent "ON" resistances, which can cause small delays in charging internal capacitances. Increasing the supply voltage reduces this "ON" resistance, which improves the accuracy of the filter in high-frequency applications. The maximum practical center frequency improves by roughly 10% to 20% when the supply voltage increases from 5V to 10V.

Dynamic range is also affected by supply voltage. Both the noise level and the maximum signal voltage increase as supply voltage increases, but the maximum signal voltage increases more rapidly with supply voltage. Thus, the dynamic range is greater with higher supply voltages. It is therefore recommended that the supply voltage be kept near the maximum operating voltage when dynamic range and/or high-frequency performance are important.

As with all switched-capacitor filters, each of the LMF120's power supply pins should be bypassed with a minimum of 0.1 μF located close to the chip.

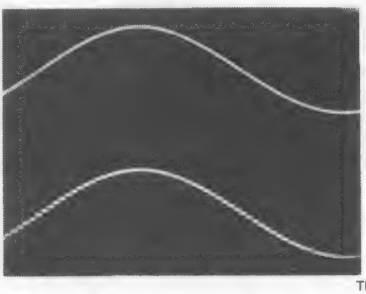

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FIGURE 2. Switched-Capacitor Filter Output Waveform. Note the sampling "steps".

SAMPLED-DATA SYSTEM CONSIDERATIONS

Output Steps

Because the LMF120 uses switched-capacitor techniques, its performance differs in several ways from non-sampled (continuous) circuits. The analog signal at any input is sampled during each filter clock cycle, and since the output voltage can change only once every clock cycle, the result is a discontinuous output signal. The output signal takes the form of a series of voltage "steps", as shown in *Figure 2*. The steps are smaller when the ratio of clock frequency to signal frequency is larger.

Aliasing

Another important characteristic of sampled-data systems is their effect on signals at frequencies greater than one-half the sampling frequency, f_S. (The LMF120's sampling frequency is the same as the filter clock frequency). If a signal with a frequency greater than one-half the sampling frequency is applied to the input of a sampled-data system, it will be "reflected" to a frequency less than one-half the sampling frequency. Thus, an input signal whose frequency is $f_s/2 + 10$ Hz will cause the system to respond as though the input frequency was $f_s/2 - 10$ Hz. This phenomenon is known as "aliasing". Aliasing can be reduced or eliminated by limiting the input signal spectrum to less than $f_s/2$. In some cases, it may be necessary to use a bandwidth-limiting filter (often a simple passive RC low-pass) between the signal source and the switched-capacitor filter's input.

Clock Frequency Limitations

The performance characteristics of a switched-capacitor filter depend on the switching (clock) frequency. At very low clock frequencies (below 10 Hz), the time between clock cycles is relatively long, and small parasitic leakage currents cause the internal capacitors to discharge sufficiently to affect the filter's offset voltage and gain. This effect becomes more pronounced at elevated operating temperatures.

At higher clock frequencies, performance deviations are primarily due to the reduced time available for the internal operational amplifiers to settle. For this reason, when the filter clock is externally generated (clock divider unused), the clock waveform's duty cycle should be as close to 50% as possible, especially at high clock frequencies.

Offset Voltage

Switched-capacitor filters often have higher offset voltages than non-sampling filters with similar topologies. This is due to charge injection from the MOS switches into the sampling and integrating capacitors. The LMF120 is built using National's LMCMOS™ process for linear CMOS circuits, and has far lower input offset voltage than most other switched-capacitor filters. Typical offset voltage for an LMF120 filter will be in the 20 mV to 400 mV range, with the actual value being strongly dependent on the type of filter response being realized and the number of cascaded biquad stages needed to achieve that particular response.

Noise

Switched-capacitor filters have two kinds of noise at their outputs. There is a random, "thermal" noise component whose level is typically on the order of 250 μV. The actual value depends on the specific filter being implemented. The other kind of noise is digital clock feedthrough. This will have an amplitude in the vicinity of 10 mV rms. In some applications, the clock noise frequency is so high compared to the signal frequency that it is unimportant. In other cases,

Applications Information (Continued)

clock noise may have to be removed from the output signal with, for example, a passive low-pass filter at the LMF120's output.

Input Impedance

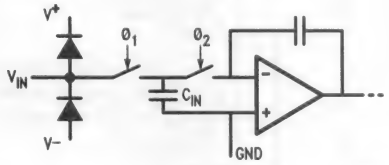
The LMF120's input pins may be connected to the sample-and-hold circuits or directly to biquad filter sections, depending on system requirements. The sample-and-hold input circuits, shown in the block diagram, are normally used only in filter implementations that require input signals (which are normally continuous) to be combined with sampled signals, as in notch and high-pass designs. Sampling the input before combining it with a sampled filter output makes the overall filter response more accurate.

During the first half of a clock cycle, the θ_1 switch closes, charging C_{IN} to the input voltage V_{IN}. During the second half-cycle, the θ_2 switch closes, and the charge on C_{IN} is transferred to the feedback capacitor. At frequencies well below the clock frequency, the input impedance approximates a resistor whose value is

$$R_{IN} = \frac{1}{C_{IN} f_{CLK}}.$$

At any sample/hold input, C_{IN} is nominally 0.5 pF. For a worst-case calculation of effective R_{IN}, assume C_{IN} = 0.5 pF and f_{CLK} = 1.5 MHz. Thus,

$$R_{in}(\text{min}) = \frac{1}{0.75 \times 10^{-6}} = 1.33\ \text{M}\Omega\,.$$



TL/F/10353–7

FIGURE 3. The inputs to the sample-and-hold circuits consist of diodes, switches, and capacitors. The input impedance has a "resistive" component that depends on the clock frequency, and a capacitive component from the protection diodes.

At the maximum clock frequency of 1.5 MHz, the lowest typical value for the effective R_{IN} the V_{IN1} input is therefore 1.33 MΩ. Note that R_{IN} increases as f_{CLK} decreases, so the input impedance will always be greater than or equal to this value. In addition to this "resistive" input impedance, the input protection diodes and the package contribute a total of about 5 pF of capacitance from the input pin to ground.

When the input pins are connected directly to a biquad section, the input impedance can be either a "pure" capacitance to ground, or a "resistive" switched-capacitor network with characteristics similar to these of the sample-and-hold circuits. As *Figure 1* shows, the capacitors at the inputs of the biquads do not have fixed values. They are typically around 1 pF to 2 pF, but can be as large as 8 pF in some designs.

Typical Applications

Third-Octave Analyzer Filter

Figure 4 is a block diagram of one version of the LMF120. The LMF120-TPQ contains three fourth-order Chebyshev bandpass filters. The center frequencies are spaced 1/3 octave apart. This circuit is intended to be used in "real time" audio spectrum analysis applications. *Figure 5* shows the computer-simulated magnitude versus frequency curves for the LMF120-TPQ. These curves meet ANSI specifications for Type E, Class II, Third-Octave filters. The center frequencies of the LMF120-TPQ's three filters are located at $f_{CLK}/50$, $f_{CLK}/62.5$, and $f_{CLK}/80$, so that by using several LMF120-TPQs with clock frequencies separated by a factor

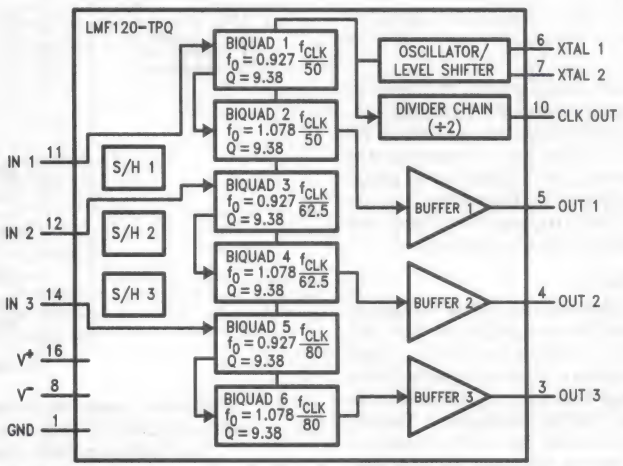

LMF120-TPQ
BIQUAD 1 f_{CLK} $f_0 = 0.927 \frac{f_{CLK}}{50}$ Q = 9.38
BIQUAD 2 f_{CLK} $f_0 = 1.078 \frac{f_{CLK}}{50}$ Q = 9.38
BIQUAD 3 f_{CLK} $f_0 = 0.927 \frac{f_{CLK}}{62.5}$ Q = 9.38
BIQUAD 4 f_{CLK} $f_0 = 1.078 \frac{f_{CLK}}{62.5}$ Q = 9.38
BIQUAD 5 f_{CLK} $f_0 = 0.927 \frac{f_{CLK}}{80}$ Q = 9.38
BIQUAD 6 f_{CLK} $f_0 = 1.078 \frac{f_{CLK}}{80}$ Q = 9.38
OSCILLATOR/ LEVEL SHIFTER
DIVIDER CHAIN (÷2)
BUFFER 1
BUFFER 2
BUFFER 3
S/H 1
S/H 2
S/H 3
IN 1 11
IN 2 12
IN 3 14
V+ 16
V− 8
GND 1
6 XTAL 1
7 XTAL 2
10 CLK OUT
5 OUT 1
4 OUT 2
3 OUT 3

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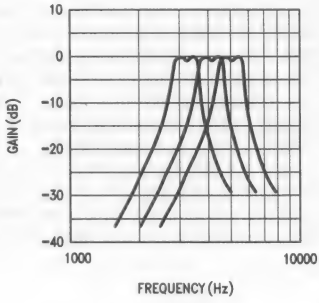
FIGURE 4. Block Diagram of LMF120-TPQ showing internal connections. Note that the input sample-and-holds are not used in this version of the LMF120. The clock output frequency is one-half of the clock input frequency.

Typical Applications (Continued)

of 2n, a complex audio program can be analyzed for frequency content over a range of several octaves. To facilitate this, the CLK OUT pin of the LMF120-TPQ supplies an output clock signal whose frequency is ½ that of the incoming clock frequency. Therefore, a single internal or external clock oscillator can provide the clock reference for all of the 30 filters in a complete audio real time analyzer.

The circuit shown in *Figure 6* uses the LMF120-TPQ to implement a ⅓-octave filter set for use in "real time" audio program analyzers. Ten LMF120-TPQs can provide all of the filtering for the full audio frequency range.

The upper LMF120 handles the highest octave, with center frequencies of 20 kHz, 16 kHz, and 12.6 kHz. It also contains the 1 MHz master clock oscillator for the entire system. Its Clock Out pin provides a 500 kHz clock for the second LMF120, which supplies 250 kHz to the third LMF120, and so on.



TL/H/10353–9

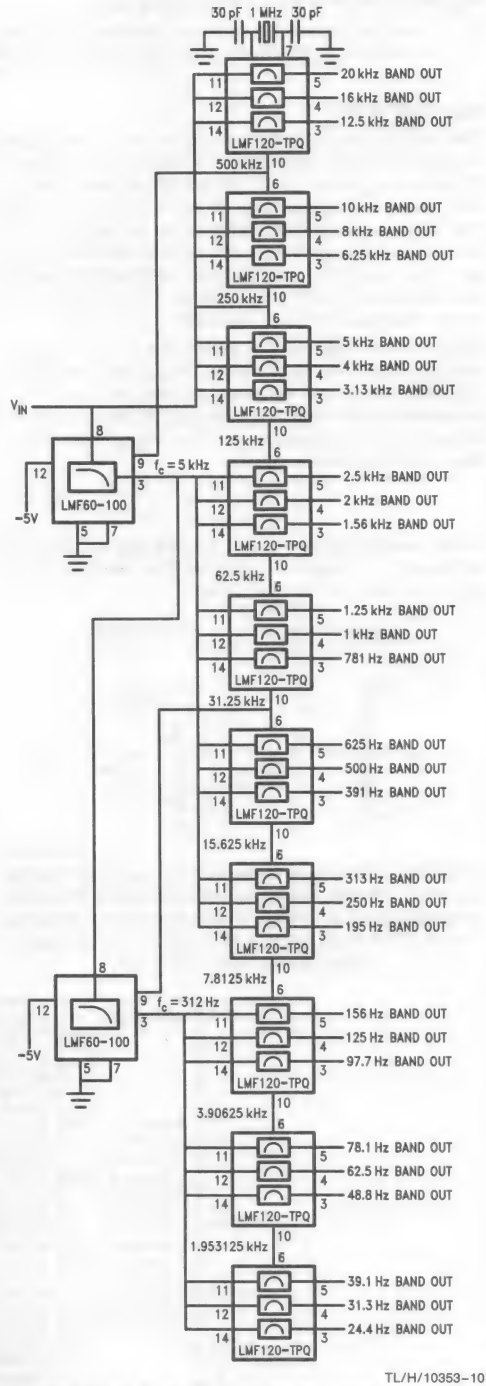
FIGURE 5. Response curves for the three filters in the LMF120-TPQ. The clock frequency is 250 kHz.

If the audio input signal were applied to all of the LMF120-TPQ input pins, aliasing might occur in the lower frequency filters due to audio components near their clock frequencies (e.g., an input signal component near 1.8 kHz will produce an output from one of the filters in the LMF120 that handles the lowest octave). This problem is solved by using two LMF60-100 6th-order Butterworth low-pass filters as anti-aliasing filters. One LMF60-100 is placed ahead of the three lowest-frequency LMF120TPQs and is clocked with the 31.25 kHz clock signal. The other LMF60-100 is ahead of the next four LMF120-TPQs and the first LMF60-100. Its clock frequency is 500 kHz.

The internal sample-and-hold circuits are not connected to the LMF120-TPQ's input pins; instead, the inputs are connected directly to C6 of three of the biquads (see *Figure 4*). C6 is 1.2 pF in the LMF120-TPQ, so the input impedance at each input of the chip handling the highest octave will be 833 kΩ. The input impedances of the filters in the next octave will be twice this, or 1.667 MΩ, and so on. Each filter will also have 5 pF of additional capacitance to ground.

12th-Order Elliptic Low-Pass

With the internal biquads connected as shown in *Figure 7*, the LMF120 functions as a 12th-order elliptic low-pass filter with 0.4 dB passband ripple. The filter's extremely rapid cutoff slope is useful in applications such as anti-aliasing filters, where unwanted signals may exist at frequencies just above those of the desired signals. Two curves of gain vs frequency are included—*Figure 8* shows the filter's overall response, and *Figure 9* shows the passband response with much higher resolution.



TL/H/10353–10

FIGURE 6. Audio "Real-Time" analyzer filter set using LMF120-TPQ one-third octave filters. The LMF60s provide anti-alias filtering. Power supplies (not shown) are ±5V and should be bypassed with 0.1 μF at each supply pin.

Typical Applications (Continued)

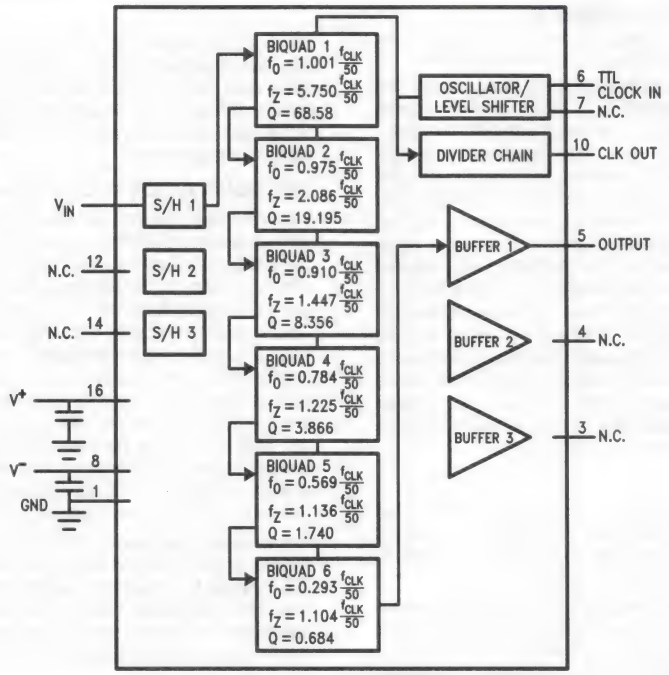


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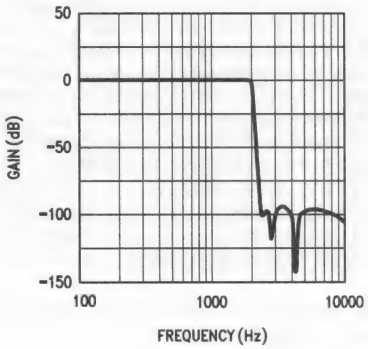
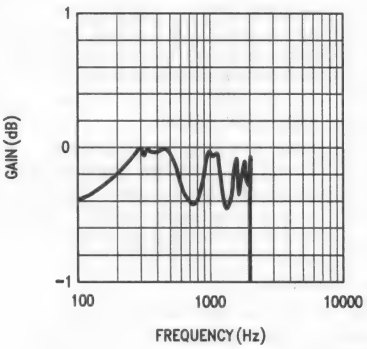
FIGURE 7. 12th-Order Elliptic Low-Pass Filter



TL/H/10353–12

FIGURE 8. Computer-simulated LMF120 12th-Order Elliptic Low-Pass Response. The clock frequency for the curve shown here is 100 kHz, and the clock-to-center-frequency ratio is 50:1.



TL/H/10353–13

FIGURE 9. Computer-simulated LMF120 12th-Order Elliptic Low-Pass Response. This curve covers the same frequency range as the one in *Figure 8*, but increased resolution shows the passband ripple more clearly.

Semi-Custom Filter Development Procedure

Note: Please contact the nearest National Semiconductor Sales Office for information on LMF120 semi-custom filter development costs.

Developing a new switched-capacitor filter using the LMF120 is relatively simple. First, define the performance requirements for the filter(s) in terms of pole and zero locations, transfer functions, or frequency/attenuation specifications, whichever is most convenient. The worksheet in the back of this data sheet may be used for this purpose. National Semiconductor will determine whether the application's performance requirements can be met with a semi-custom proprietary version of the LMF120. If the required filter is feasible, computer simulations of the filter's performance will be provided. If the performance is satisfactory, test frequencies and performance limits will be chosen and the custom metal mask will be produced and prototype devices will be manufactured. The prototyping stage generally takes from eight to twelve weeks. After prototypes have been built, tested, and approved, production can begin. (See the pre-production activity flow in *Figure 10*).

Feasibility

The first step in developing a custom filter based on the LMF120 is to determine whether an LMF120 can indeed realize the desired filter response. To this end, it is helpful to understand the limitations of the circuit.

The center or cutoff frequency (f_0) of the filter is one limitation. As indicated in the table of Filter Electrical Characteristics, this can typically range from a low of 0.1 Hz to a high of 100 kHz. These numbers, however, are given as guidelines only. The actual frequency limits will depend on the specific characteristics of the filter being developed. For example, if the desired filter must have a very fast attenuation slope beyond the cutoff frequency, the maximum cutoff frequency may be significantly less than 100 kHz. As a general rule, filters with gentler slopes can have cutoff frequencies as high as 100 kHz, while very fast rolloffs may be limited to corner frequencies below 20 kHz.

Filter Q is another parameter whose acceptable range is strongly dependent on the desired characteristics. Higher values of Q are more difficult to achieve with high center or corner frequencies. A useful figure of merit is the product of Q and F_0. If this product is less than 1 MHz and Q is less than 100 for each biquad filter section, it should be achievable with the LMF120.

Filter order is obviously an important specification. If the desired filter response requires a 13th-order filter, it can't be fully implemented by a single LMF120, which can provide up to 12 poles of filtering.

As discussed earier in this data sheet, the LMF120's offset voltage will generally be in the tens of millivolts, and will be dependent on the kind of transfer function the filter is intended to realize. It is important to ensure that the application's requirements are compatible with the LMF120's offset voltage characteristics.

THE DESIGN AUTOMATION SYSTEM

National Semiconductor customizes the LMF120 to a specific application by generating a metal mask that provides the interconnections between the internal circuit blocks and programs them for the required characteristics. The mask is generated using National's proprietary filter CAD software. This software computes the optimum capacitor values for each of the six switched-capacitor biquad filter sections to ensure close conformance to the target requirements. It also optimizes the design for high signal-to-noise ratio, and then analyzes the design, taking into account all second-order effects, such as parasitic capacitances, switch "ON" resistance, and the finite gain-bandwidth products of the operational amplifiers. The final design analysis is then returned for verification and approval.

Actual metal mask generation begins once the design and the test frequencies and limits have been approved. National's in-house CAD system is used to facilitate mask generation. The new metal mask is then used to complete the fabrication of the final silicon. The design automation system ensures fast and accurate results on the first run.

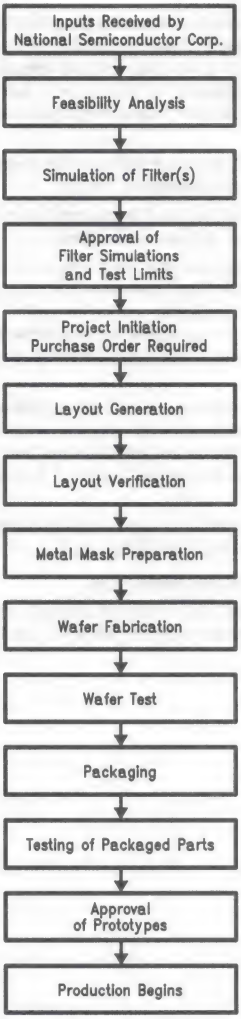
The Test Procedure

When the IC is in production, its performance must be verified by automated testing. Some of the tests will be common to all versions of the LMF120: logic levels and logic input current for example. Other tests will be for parameters that are specific to a particular metal mask. These consist of total supply current, DC offset voltages, signal swing, and several frequency/gain (or attenuation) test points for each filter. The frequencies and test limits will be tailored to the specific application requirements for the filter(s).

National will provide information on the typical behavior of the filter(s) for those parameters that are not tested or guaranteed by design, such as clock feedthrough and output noise. This information will be returned with the prototype parts.

Some special test requirements can be accommodated; these will be evaluated on request.

Semi-Custom Filter Development Procedure (Continued)



TL/H/10353–14

FIGURE 10. Pre-Production Activity Flow

LMF120 Filter Worksheet Instructions

Use the following instructions for completing the attached Filter Worksheet. Return one completed worksheet for each LMF120 device (maximum of three filters per LMF120) to your local National Semiconductor sales office. If you require more worksheets you may photocopy this one.

1. Supply Voltage:

Specify your system supply voltage requirements. The total supply voltage ($V^+ - V^-$) can be anywhere from 4V to 14V. Higher voltages are advantageous when dynamic range and maximum operating frequency are critical concerns.

2. Total Number of Filters:

The LMF120 may consist of one, two or three independent filters. Specify the total number of filters for this LMF120 design.

3. Input Clock Frequency:

The maximum clock frequency is 4 MHz. Specify the crystal frequency if you plan to use the internal crystal oscillator. Two external capacitors and one crystal are required for the crystal oscillator.

4. Filter Clock Frequency:

This is the frequency at which the filter will be clocked. There are many factors to be considered in the choice of this frequency. Operation at the highest possible clock frequency reduces aliasing in the signal band, and reduces the need for pre- and/or post-filtering. However, there are certain factors that limit the maximum frequency. These include finite gain-bandwidth of the op-amps and finite on-resistances of internal switches. On the other hand, using slow clock frequencies enables the filter to operate at lower supply currents and to save power on applications requiring low-power operation. The maximum clock frequency for the LMF120's internal biquads is 1.5 MHz, so the internal clock frequency divider must be used to reduce this frequency if the clock frequency at the LMF120's clock input pin is greater than 1.5 MHz. Additionally, the filter clock frequency must also be at least ten times higher (and preferably 50 to 100 times higher) than the highest pole or zero in the filter structure.

Example: Determine the filter clock frequency for a BANDPASS filter with center at 1 kHz. The system clock (input clock) is 3.5 MHz.

Solution: Since the input clock is higher than 1.5 MHz it must be divided down internally. Dividing by 32 gives a filter clock frequency of 109.38 kHz. Therefore, the clock-to-center frequency ratio is 109380/1000 = 109.38. This is close to the 50:1 to 100:1 range of clock-to-center-frequency ratios that generally gives the best results.

5. Clock Output Frequency:

This is an optional output that may be used to supply a clock frequency anywhere else in the application system. This output is subject to the following constraints:

$f_{CLKOUT} = f_{CLKIN}/2^n$ for n = 0, 1, . . . 8. Specify N/A if this output is not to be used.

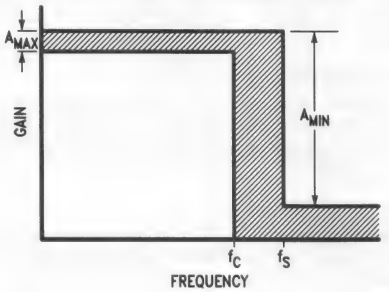
6. Input Clock Level:

CMOS or TTL input levels may be specified for 0V–5V, ±5V or 0V–10V power supplies. For non-standard supplies, only CMOS input levels may be specified.

7. Filter Descriptions:

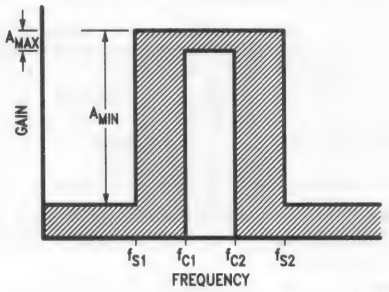
Use this space to describe the filter(s) by transfer functions, band diagrams, pole-zero locations, or f_0 and Q values for the individual biquads. Pole-zero locations or f_0 and Q values are preferred, but the filters may be described in any of the ways mentioned above. Examples of appropriate band diagrams are shown in *Figures 11* and *12*. f_C is the cutoff frequency of the passband and f_S is the frequency that defines the beginning of the stopband. A_{MAX} is the maximum acceptable passband gain variation. A_{MIN} is the minimum acceptable stopband attenuation.

LMF120 Filter Worksheet Instructions (Continued)



TL/H/10353–15

FIGURE 11. Format of a band diagram for a low-pass filter. The amplitude response requirements are specified by A_{MAX}, A_{MIN}, f_C and f_S.



TL/H/10353–16

FIGURE 12. Format of a band diagram for a bandpass filter. The filter's amplitude response requirements are specified by A_{MIN}, A_{MAX}, f_{C1}, f_{C2}, f_{S1} and f_{S2}.

Test frequencies for each filter should be specified with the following in mind:

A. Test frequencies between 100 Hz and 8 kHz: Digital Signal Processing techniques are used in the test procedure. This produces the best accuracy and allows the measurement of both amplitude and phase response at the test frequencies. The customer may choose between the following alternatives:

1. 7 test frequencies; each frequency is a multiple of 10 Hz with a minimum difference of 10 Hz.
2. 15 test frequencies; each frequency is a multiple of 10 Hz with a minimum difference of 20 Hz.

In the DSP test procedure, all of the test frequencies are applied to the filter simultaneously. The output energy available at any given frequency will be less with 15 test frequencies than with 7 test frequencies; therefore the test will be more accurate with 7 test frequencies than with 15 test frequencies.

B. Test frequencies above 8 kHz will require a voltmeter test method, which can measure only the amplitude response. The only constraint on the voltmeter method is that the test frequencies must be above 1 kHz. 7 frequencies can be tested.

Any special requirements will be considered separately, and may be included with the Worksheet.

8. APPLICATION INFORMATION:

Describe the application, the end product, and the most important performance characteristics for the filter in this application.

LMF120 Filter Worksheet

Engineering Contact _____

Phone _____

Company Name _____

Address _____

_____

_____

1) Supply Voltage _____

2) Total Number of Filters _____

3) Input Clock Frequency (4 MHz Maximum) _____

4) Filter Clock (1.5 MHz Maximum) _____

5) Clock Output Frequency _____

6) Input Clock Logic Levels (TTL or CMOS) _____

7) Filter Descriptions:

Please use the space below to define your filter(s). Note that the total sum of the poles or zeros for all three filters must not exceed twelve.

Filter #1

Filter Order _____

Use the space below to write the transfer function or pole/zero locations, or plot a detailed band diagram. Use a separate page if more space is needed.

Filter #1 Test Frequencies _____

_____

_____

_____

_____

_____

_____

Filter #2

Filter Order _____

Use the space below to write the transfer function or pole/zero locations, or plot a detailed band diagram. Use a separate page if more space is needed.

Filter #2 Test Frequencies _____

_____

_____

_____

_____

_____

_____

Filter #3

Filter Order _____

Use the space below to write the transfer function or pole/zero locations, or plot a detailed band diagram. Use a separate page if more space is needed.

Filter #3 Test Frequencies _____

_____

_____

_____

_____

_____

_____

8) Application Information: Please describe in detail the application for the LMF120 filters in your system.

a) End Product: _____

_____

b) Projected Volume per Year: _____

c) List the most important performance requirements for the filters in your application (i.e., Dynamic Range >50 dB, etc.)

National Semiconductor

MF4 4th Order Switched Capacitor Butterworth Lowpass Filter

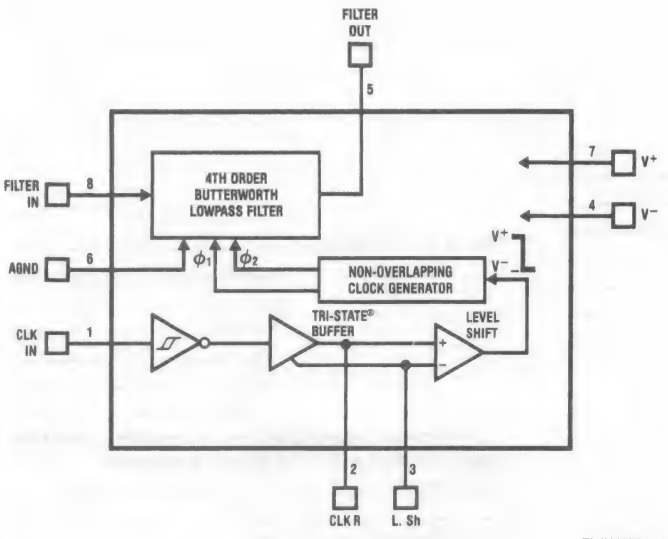
General Description

The MF4 is a versatile, easy to use, precision 4th order Butterworth low-pass filter. Switched-capacitor techniques eliminate external component requirements and allow a clock-tunable cutoff frequency. The ratio of the clock frequency to the low-pass cutoff frequency is internally set to 50 to 1 (MF4-50) or 100 to 1 (MF4-100). A Schmitt trigger clock input stage allows two clocking options, either self-clocking (via an external resistor and capacitor) for stand-alone applications, or for tighter cutoff frequency control an external TTL or CMOS logic compatible clock can be applied. The maximally flat passband frequency response together with a DC gain of 1 V/V allows cascading MF4 sections together for higher order filtering.

Features

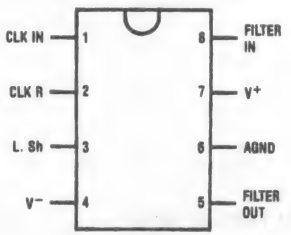
- Low Cost
- Easy to use
- 8-pin mini-DIP or 14-pin wide-body S.O.
- No external components
- 5V to 14V supply voltage
- Cutoff frequency range of 0.1 Hz to 20 kHz
- Cutoff frequency accuracy of ± 0.3% typical
- Cutoff frequency set by external clock
- Separate TTL and CMOS/Schmitt-trigger clock inputs

Block and Connection Diagrams



TL/H/5064–1

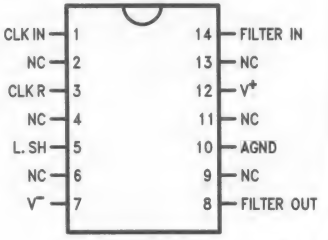
Dual-In-Line Package



TL/H/5064–2

Order Number MF4CN-50
or MF4CN-100
See NS Package Number N08E

Small-Outline
Wide-Body Package



TL/H/5064–25

Top View
Order Number MF4CWM-50
or MF4CWM-100
See NS Package Number M14B

Absolute Maximum Ratings (Notes 1, 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage ($V^+ - V^-$)	14V
Voltage At Any Pin	$V^+ + 0.2V$ $V^- - 0.2V$
Input Current at Any Pin (Note 14)	5 mA
Package Input Current (Note 14)	20 mA
Power Dissipation (Note 15)	500 mW
Storage Temperature	150°C
ESD Susceptibility (Note 13)	800 V

Soldering Information:

- N Package: 10 sec. 260°C
- SO Package: Vapor Phase (60 sec.) 215°C
  Infrared (15 sec.) 220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

Operating Ratings (Note 2)

Temperature Range	$T_{min} \le T_A \le T_{max}$
MF4CN-50, MF4CN-100	$0°C \le T_A \le 70°C$
MF4CWM-50, MF4CWM-100	$0°C \le T_A \le 70°C$
Supply Voltage ($V^+ - V^-$)	5V to 14V

Filter Electrical Characteristics

The following specifications apply for $f_{CLK} \le 250$ kHz (see Note 5) unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Parameter	Conditions	MF4-50			MF4-100			Unit
		Typical (Note 10)	Tested Limit (Note 11)	Design Limit (Note 12)	Typical (Note 10)	Tested Limit (Note 11)	Design Limit (Note 12)	
$V^+ = +5V, V^- = -5V$								
f_c, Cutoff Frequency Range (Note 3) Min				0.1			0.1	Hz
f_c, Cutoff Frequency Range (Note 3) Max				20k			10k	Hz
Supply Current	$f_{clk} = 250$ kHz	2.5	3.5	**3.5**	2.5	3.5	**3.5**	mA
Maximum Clock Feedthrough (Peak-to-Peak) Filter Output	$V_{in} = 0V$	25			25			mV
H_o, DC Gain	$R_{source} \le 2$ kΩ	0.0	±0.15	**±0.15**	0.0	±0.15	**±0.15**	dB
f_{clk}/f_c, Clock to Cutoff Frequency Ratio		49.96 ±0.3%	49.96 ±0.8%	**49.96** **±0.6%**	99.09 ±0.3%	99.09 ±1.0%	**99.09** **±0.6%**	
f_{clk}/f_c Temperature Coefficient		±15			±30			ppm/°C
Stopband Attenuation (Min)	at 2 f_c	−25.0	−24.0	**−24.0**	−25.0	−24.0	**−24.0**	dB
DC Offset Voltage		−200			−400			mV
Minimum Output Swing	$R_L = 10$ kΩ	+4.0 −4.5	+3.5 −4.0	**+3.5** **−4.0**	+4.0 −4.5	+3.5 −4.0	**+3.5** **−4.0**	V V
Output Short Circuit Current (Note 8) Source		50			50			mA
Output Short Circuit Current (Note 8) Sink		1.5			1.5			mA
Dynamic Range (Note 4)		80			82			dB
Additional Magnitude Response Test Points (Note 6) $f_{clk} = 250$ kHz	f = 6000 Hz		−7.57 ±0.27	**−7.57** **±0.27**				dB
	f = 4500 Hz		−1.44 ±0.12	**−1.44** **±0.12**				
	f = 3000 Hz					−7.21 ±0.2	**−7.21** **±0.2**	dB
	f = 2250 Hz					−1.39 ±0.1	**−1.39** **±0.1**	

Filter Electrical Characteristics

The following specifications apply for $f_{CLK} \leq 250$ kHz (see Note 5) unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$. (Continued)

Parameter	Conditions	MF4-50			MF4-100			Unit
		Typical (Note 10)	Tested Limit (Note 11)	Design Limit (Note 12)	Typical (Note 10)	Tested Limit (Note 11)	Design Limit (Note 12)	
$V^+ = +2.5V, V^- = -2.5V$								
f_c Cutoff Frequency Range (Note 3) min				0.1			0.1	Hz
f_c Cutoff Frequency Range (Note 3) max				10k			5k	Hz
Supply Current	$f_{clk} = 250$ kHz	1.5	2.25	**2.25**	1.5	2.25	**2.25**	mA
Maximum Clock Feedthrough (Peak-to-Peak) Filter Output	$V_{in} = 0V$	15			15			mV
H_o, DC Gain	$R_{source} \leq 2$ kΩ	0.0	±0.15	**±0.15**	0.0	±0.15	**±0.15**	dB
f_{clk}/f_c, Clock to Cutoff Frequency Ratio		50.07 ±0.3%	50.07 ±1.0%	**50.07** **±0.6%**	99.16 ±0.3%	99.16 ±1.0%	**99.16** **±0.6%**	
f_{CLK}/f_C Temperature Coefficient		±25			±60			ppm/°C
Stopband Attenuation (Min)	at 2 f_c	−25.0	−24.0	**−24.0**	−25.0	−24.0	**−24.0**	dB
DC Offset Voltage		−150			−300			mV
Minimum Output Swing	$R_L = 10$ kΩ	+1.5 −2.2	+1.0 −1.7	**+1.0** **−1.7**	+1.5 −2.2	+1.0 −1.7	**+1.0** **−1.7**	V V
Output Short Circuit Current (Note 8) Source		28			28			mA
Output Short Circuit Current (Note 8) Sink		0.5			0.5			mA
Dynamic Range (Note 4)		78			78			dB
Additional Magnitude Response Test Points (Note 6) ($f_c = 5$ kHz) Magnitude at	$f_{clk} = 250$ kHz f = 6000 Hz		−7.57 ±0.27	**−7.57** **±0.27**				dB
	f = 4500 Hz		−1.46 ±0.12	**−1.46** **±0.12**				dB
($f_c = 2.5$ kHz) Magnitude	f = 3000 Hz					−7.21 ±0.2	**−7.21** **±0.2**	dB
	f = 2250 Hz					−1.39 ±0.1	**−1.39** **±0.1**	dB

Logic Input-Output Characteristics

The following specifications apply for $V^- = 0V$ (see Note 7) unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Parameter		Conditions	Typical (Note 10)	Tested Limit (Note 11)	Design Limit (Note 12)	Unit
SCHMITT TRIGGER						
V_{T+}, Positive Going Threshold Voltage	Min	$V^+ = 10V$	7.0	6.1	**6.1**	V
	Max				**8.9**	V
	Min	$V^+ = 5V$	3.5	3.1	**3.1**	V
	Max			4.4	**4.4**	V

Logic Input-Output Characteristics

The following specifications apply for $V^- = 0V$ (see Note 7) unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = t_J = 25°C$. (Continued)

Parameter		Conditions		Typical (Note 10)	Tested Limit (Note 11)	Design Limit (Note 12)	Unit
SCHMITT TRIGGER (Continued)							
V_{T-}, Negative Going Threshold Voltage	Min	$V^+ = 10V$		3.0	1.3	**1.3**	V
	Max				3.8	**3.8**	
	Min	$V^+ = 5V$		1.5	0.6	**0.6**	V
	Max				1.9	**1.9**	
Hysteresis ($V_{T+} - V_{T-}$)	Min	$V^+ = 10V$		4.0	2.3	**2.3**	V
	Max				7.6	**7.6**	
	Min	$V^+ = 5V$		2.0	1.2	**1.2**	V
	Max				3.8	**3.8**	
Minimum Logical "1" Output Voltage (pin 2)		$I_0 = -10\ \mu A$	$V^+ = 10V$		9.0	**9.0**	V
			$V^+ = 5V$		4.5	**4.5**	V
Maximum Logical "0" Output Voltage (pin 2)		$I_0 = 10\ \mu A$	$V^+ = 10V$		1.0	**1.0**	V
			$V^+ = 5V$		0.5	**0.5**	V
Minimum Output Source Current (pin 2)		CLK R Shorted to Ground	$V^+ = 10V$	6.0	3.0	**3.0**	mA
			$V^+ = 5V$	1.5	0.75	**0.75**	mA
Maximum Output Sink Current (pin 2)		CLK R Shorted to V^+	$V^+ = 10V$	5.0	2.5	**2.5**	mA
			$V^+ = 5V$	1.3	0.65	**0.65**	mA
TTL CLOCK INPUT, CLK R PIN (Note 9)							
Maximum V_{IL}, Logical "0" Input Voltage				0.8			V
Minimum V_{IH}, Logical "1" Input Voltage				2.0			V
Maximum Leakage Current at CLK R Pin		L. Sh Pin at Mid-Supply		2.0			μA

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. AC and DC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are with respect to GND.

Note 3: The cutoff frequency of the filter is defined as the frequency where the magnitude response is 3.01 dB less than the DC gain of the filter.

Note 4: For $\pm 5V$ supplies the dynamic range is referenced to 2.82 Vrms (4V peak) where the wideband noise over a 20 kHz bandwidth is typically 280 μVrms for the MF4-50 and 230 μVrms for the MF4-100. For $\pm 2.5V$ supplies the dynamic range is referenced to 1.06 Vrms (1.5V peak) where the wideband noise over a 20 kHz bandwidth is typically 130 μVrms for both the MF4-50 and the MF4-100.

Note 5: The specifications for the MF4 have been given for a clock frequency (f_{CLK}) of 250 kHz or less. Above ths clock frequency the cutoff frequency begins to deviate from the specified error band of $\pm 0.6\%$ but the filter still maintains its magnitude characteristics. See Application Hints.

Note 6: Besides checking the cutoff frequency (f_c) and the stopband attenuation at 2 f_c, two additional frequencies are used to check the magnitude response of the filter. The magnitudes are referenced to a DC gain of 0.0 dB.

Note 7: For simplicity all the logic levels have been referenced to $V^- = 0V$ (except for the TTL input logic levels). The logic levels will scale accordingly for $\pm 5V$ and $\pm 2.5V$ supplies.

Note 8: The short circuit source current is measured by forcing the output that is being tested to its maximum positive voltage swing and then shorting that output to the negative supply. The short circuit sink current is measured by forcing the output that is being tested to its maximum negative voltage and then shorting that output to the positive supply. These are worst case conditions.

Note 9: The MF4 is operating with symmetrical split supplies and L. Sh is tied to ground.

Note 10: Typicals are at 25°C and represent most likely parametric norm.

Note 11: Guaranteed to National's Average Outgoing Quality Level (AOOL).

Note 12: Guaranteed, but not 100% production tested. These limits are not used to determine outgoing quality levels.

Note 13: Human body model; 100 pF discharged through a 1.5 kΩ resistor.

Note 14: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 5 mA or less. The 20 mA package input current limits the number of pins that can exceed the power supply boundaries with a 5 mA current limit to four.

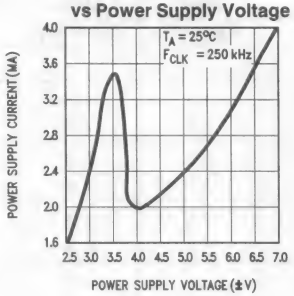
Note 15: Thermal Resistance

θ_{JA} (Junction to Ambient) N Package ..... 105°C/W.

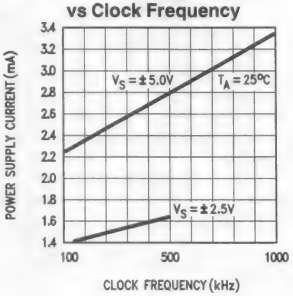
θ_{JA} M Package ..... 95°C/W.

Typical Performance Characteristics

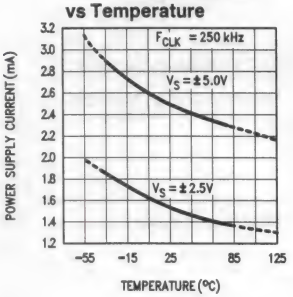
Power Supply Current vs Power Supply Voltage



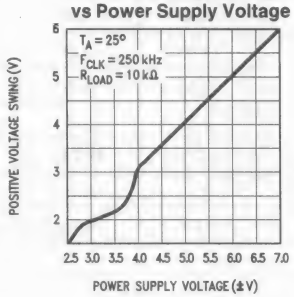
Power Supply Current vs Clock Frequency



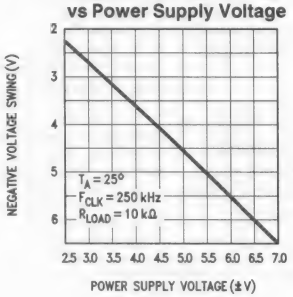
Power Supply Current vs Temperature



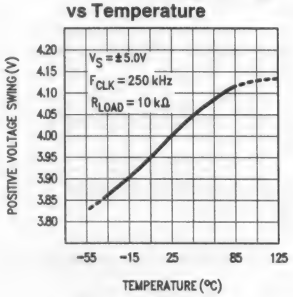
Positive Voltage Swing vs Power Supply Voltage



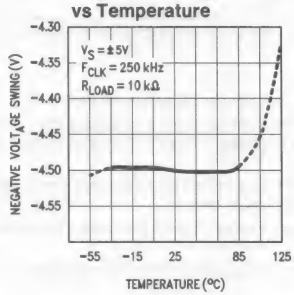
Negative Voltage Swing vs Power Supply Voltage



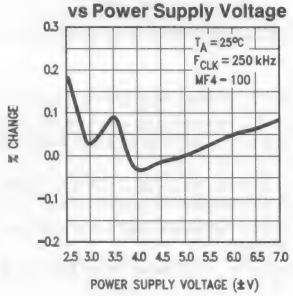
Positive Voltage Swing vs Temperature



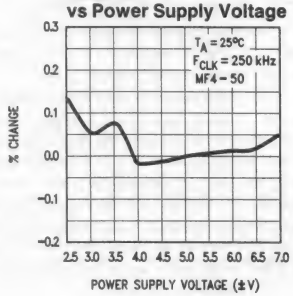
Negative Voltage Swing vs Temperature



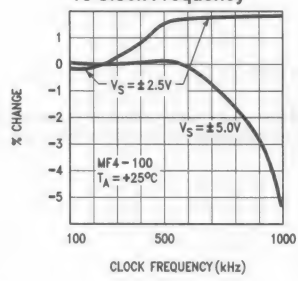
f_{CLK}/f_C Deviation vs Power Supply Voltage



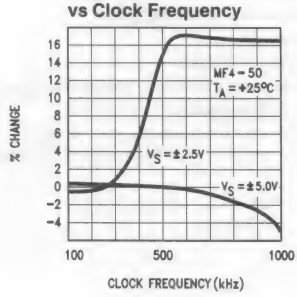
f_{CLK}/f_C Deviation vs Power Supply Voltage



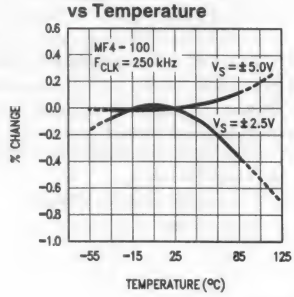
f_{CLK}/f_C Deviation vs Clock Frequency



f_{CLK}/f_C Deviation vs Clock Frequency



f_{CLK}/f_C Deviation vs Temperature



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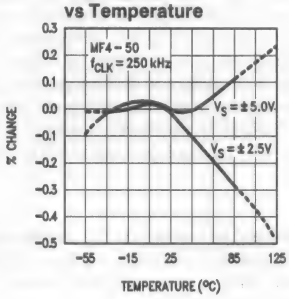
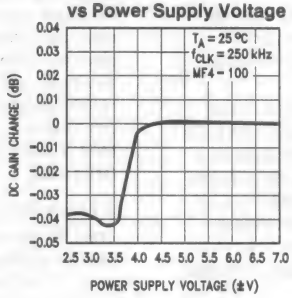
Typical Performance Characteristics (Continued)

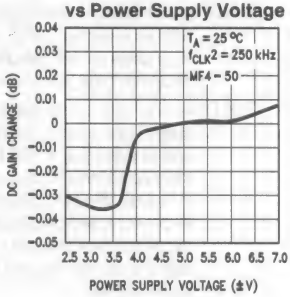
f_{CLK}/f_c Deviation vs Temperature



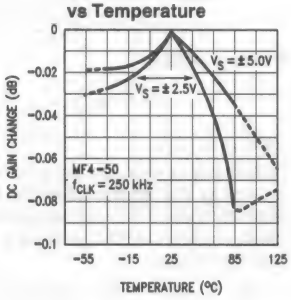
DC Gain Deviation vs Power Supply Voltage



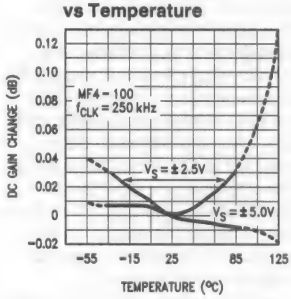
DC Gain Deviation vs Power Supply Voltage



DC Gain Deviation vs Temperature



DC Gain Deviation vs Temperature



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Pin Descriptions

(Numbers in ( ) are for 14-pin package.)

Pin #	Pin Name	Function
1 (1)	CLK IN	A CMOS Schmitt-trigger input to be used with an external CMOS logic level clock. Also used for self clocking Schmitt-trigger oscillator (see section 1.1).
2 (3)	CLK R	A TTL logic level clock input when in split supply operation (±2.5V to ±7V) with L. Sh tied to system ground. This pin becomes a low impedance output when L. Sh is tied to V^-. Also used in conjunction with the CLK IN pin for a self clocking Schmitt-trigger oscillator (see section 1.1). The TTL input signal must not exceed the supply voltages by more than 0.2V.
3 (5)	L. Sh	Level shift pin; selects the logic threshold levels for the clock. When tied to V^- it enables an internal tri-state buffer stage between the Schmitt trigger and the internal clock level shift stage thus enabling the CLK IN Schmitt-trigger input and making the CLK R pin a low impedance output. When the voltage level at this input exceeds 25% $(V^+ - V^-) + V^-$ the internal tri-state buffer is disabled allowing the CLK R pin to become the clock input for the internal clock level-shift stage. The CLK R threshold level is now 2V above the voltage on the L. Sh pin. The CLK R pin will be compatible with TTL logic levels when the MF4 is operated on split supplies with the L. Sh pin connected to system ground.
5 (8)	FILTER OUT	The output of the low-pass filter. It will typically sink 0.9 mA and source 3 mA and swing to within 1V of each supply rail.
6 (10)	AGND	The analog ground pin. This pin sets the DC bias level for the filter section and must be tied to the system ground for split supply operation or to mid-supply for single supply operation (see section 1.2). When tied to mid-supply this pin should be well bypassed.
7, 4 (7, 12)	V^+, V^-	The positive and negative supply pins. The total power supply range is 5V to 14V. Decoupling these pins with 0.1 μF capacitors is highly recommended.
8 (14)	FILTER IN	The input to the low-pass filter. To minimize gain errors the source impedance that drives this input should be less than 2K (see section 1.3 of the Application Hints). For single supply operation the input signal must be biased to mid-supply or AC coupled through a capacitor.

1.0 MF4 Application Hints

The MF4 is a non-inverting unity gain low-pass fourth-order Butterworth switched-capacitor filter. The switched-capacitor topology makes the cutoff frequency (where the gain drops 3.01 dB below the DC gain) a direct ratio (100:1 or 50:1) of the clock frequency supplied to the filter. Internal integrator time constants set the filter's cutoff frequency. The resistive element of these integrators is actually a capacitor which is "switched" at the clock frequency (for a detailed discussion see Input Impedance Section). Varying the clock frequency changes the value of this resistive element and thus the time constant of the integrators. The clock-to-cutoff-frequency ratio ($f_{CLK}f_c$) is set by the ratio of the input and feedback capacitors in the integrators. The higher the clock-to-cutoff-frequency ratio the closer this approximation is to the theoretical Butterworth response. The MF4 is available in f_{CLK}/f_c ratios of 50:1 (MF4-50) or 100:1 (MF4-100).

1.1 CLOCK INPUTS

The MF4 has a Schmitt-trigger inverting buffer which can be used to construct a simple R/C oscillator. Pin 3 is connected to V^- which makes Pin 2 a low impedance output. The oscillator's frequency is nominally

$$f_{CLK} = \frac{1}{RC \ln\left[\left(\frac{V_{CC} - V_{T-}}{V_{CC} - V_{T+}}\right)\left(\frac{V_{T+}}{V_{T-}}\right)\right]} \quad (1)$$

which, is typically

$$f_{CLK} \cong \frac{1}{1.69\ RC} \quad (1a)$$

for $V_{CC} = 10V$.

Note that f_{CLK} is dependent on the buffer's threshold levels as well as the resistor/capacitor tolerance (see *Figure 1*). Schmitt-trigger threshold voltage levels can change significantly causing the R/C oscillator's frequency to vary greatly from part to part.

Where accurate cutoff frequency is required, an external clock can be used to drive the CLK R input of the MF4. This input is TTL logic level compatible and also presents a very light load to the external clock source (~2 μA). With split supplies and the level shift (L. Sh) tied to system ground, the logic level is about 2V. (See the Pin Description for L. Sh).

1.2 POWER SUPPLY

The MF4 can be powered from a single supply or split supplies. The split supply mode shown in *Figure 2* is the most flexible and easiest to implement. Supply voltages of ±5V to ±7V enable the use of TTL or CMOS clock logic levels. *Figure 3* shows AGND resistor-biased to $V^+/2$ for single supply operation. In this mode only CMOS clock logic levels can be used, and input signals should be capacitor-coupled or biased near mid-supply.

1.3 INPUT IMPEDANCE

The MF4 low-pass filter input (FILTER IN) is not a high impedance buffer input. This input is a switched-capacitor resistor equivalent, and its effective impedance is inversely proportional to the clock frequency. The equivalent circuit of the filter's input can be seen in *Figure 4*. The input capacitor charges to V_{in} during the first half of the clock period; during the second half the charge is transferred to the feedback capacitor. The total transfer of charge in one clock cycle is therefore $Q = C_{in}V_{in}$, and since current is defined as the flow of charge per unit time, the average input current becomes

$$I_{in} = Q/T$$

1.0 MF4 Application Hints (Continued)

(where T equals one clock period) or

$$I_{in} = \frac{C_{in}V_{in}}{T} = C_{in}V_{in}f_{CLK}$$

The equivalent input resistor (R_{in}) then can be expressed as

$$R_{in} = \frac{V_{in}}{I_{in}} = \frac{1}{C_{in}f_{CLK}}$$

The input capacitor is 2 pF for the MF4-50 and 1 pF for the MF4-100, so for the MF4-100

$$R_{in} = \frac{1 \times 10^{12}}{f_{CLK}} = \frac{1 \times 10^{12}}{f_c \times 100} = \frac{1 \times 10^{10}}{f_c}$$

and

$$R_{in} = \frac{5 \times 10^{11}}{f_{CLK}} = \frac{5 \times 10^{11}}{f_c \times 50} = \frac{1 \times 10^{10}}{f_c}$$

for the MF4-50. The above equation shows that for a given cutoff frequency (f_c), the input resistance of the MF4-50 is the same as that of the MF4-100. The higher the clock-to-cutoff-frequency ratio, the greater equivalent input resistance for a given clock frequency.

This input resistance will form a voltage divider with the source impedance (R_{source}). Since R_{in} is inversely proportional to the cutoff frequency, operation at higher cutoff frequencies will be more likely to load the input signal which would appear as an overall decrease in gain to the output of the filter. Since the filter's ideal gain is unity, the overall gain is given by:

$$A_v = \frac{R_{in}}{R_{in} + R_{source}}$$

If the MF4-50 or the MF-100 were set up for a cutoff frequency of 10 kHz the input impedance would be:

$$R_{in} = \frac{1 \times 10^{10}}{10\ \text{kHz}} = 1\ \text{M}\Omega$$

In this example with a source impedance of 10K the overall gain, if the MF4 had an ideal gain of 1 or 0 dB, would be:

$$A_v = \frac{1\ \text{M}\Omega}{10\ \text{k}\Omega + 1\ \text{M}\Omega} = 0.99009 \text{ or } -0.086\ \text{dB}$$

Since the maximum overall gain error for the MF4 is ±0.15 dB with $R_s \leq 2\ \text{k}\Omega$ the actual gain error for this case would be +0.06 dB to −0.24 dB.

1.4 CUTOFF FREQUENCY RANGE

The filter's cutoff frequency (f_c) has a lower limit due to leakage currents through the internal switches draining the charge stored on the capacitors. At lower clock frequencies these leakage currents can cause millivolts of error, for example:

$$f_{CLK} = 100\ \text{Hz}, I_{leakage} = 1\ \text{pA}, C = 1\ \text{pF}$$

$$V = \frac{1\ \text{pA}}{1\ \text{pF (100 Hz)}} = 10\ \text{mV}$$

The propagation delay in the logic and the settling time required to acquire a new voltage level on the capacitors limit the filter's accuracy at high clock frequencies. The amplitude characteristic on ±5V supplies will typically stay flat until f_{CLK} exceeds 750 kHz and then peak at about 0.5 dB at the corner frequency with a 1 MHz clock. As supply voltage drops to ± 2.5V, a shift in the f_{CLK}/f_c ratio occurs which will become noticeable when the clock frequency exceeds 250 kHz. The response of the MF4 is still a good approximation of the ideal Butterworth low-pass characteristic shown in *Figure 5*.

2.0 Designing With The MF4

Given any low-pass filter specification, two equations will come in handy in trying to determine whether the MF4 will do the job. The first equation determines the order of the low-pass filter required to meet a given response specification:

$$n = \frac{\log\left[(10^{0.1A_{min}} - 1)/(10^{0.1A_{max}} - 1)\right]}{2\log(f_s/f_b)} \quad (2)$$

where n is the order of the filter, A_{min} is the minimum stopband attenuation (in dB) desired at frequency f_s, and A_{max} is the passband ripple or attenuation (in dB) at cutoff frequency f_b. If the result of this equation is greater than 4, more than a single MF4 is required.

The attenuation at any frequency can be found by the following equation:

$$\text{Attn}(f) = 10\log\left[1 + (10^{0.1A_{max}} - 1)(f/f_b)^{2n}\right]\ \text{dB} \quad (3)$$

where n = 4 for the MF4.

2.1 A LOW-PASS DESIGN EXAMPLE

Suppose the amplitude response specification in *Figure 6* is given. Can the MF4 be used? The order of the Butterworth approximation will have to be determined using (1):

A_{min} = 18 dB, A_{max} = 1.0 dB, f_s = 2 kHz, and f_b = 1 kHz

$$n = \frac{\log\left[(10^{1.8} - 1)/(10^{0.1} - 1)\right]}{2\log(2)} = 3.95$$

Since n can only take on integer values, n = 4. Therefore the MF4 can be used. In general, if n is 4 or less a single MF4 stage can be utilized.

Likewise, the attenuation at f_s can be found using (3) with the above values and n = 4:

$$\text{Attn (2 kHz)} = 10\log\left[1 + 10^{0.1} - 1)\,(2\ \text{kHz}/1\ \text{kHz})^8\right] = 18.28\ \text{dB}$$

This result also meets the design specification given in *Figure 6* again verifying that a single MF4 section will be adequate.

Since the MF4's cutoff frequency (f_c), which corresponds to a gain attenuation of −3.01 dB, was not specified in this example, it needs to be calculated. Solving equation 3 where $f = f_c$ as follows:

$$f_c = f_b\left[\frac{(10^{0.1(3.01\ \text{dB})} - 1}{(10^{0.1A_{max}} - 1)}\right]^{1/(2n)}$$

$$= 1\ \text{kHz}\left[\frac{10^{0.301} - 1}{10^{0.1} - 1}\right]^{1/8}$$

$$= 1.184\ \text{kHz}$$

where $f_c = f_{CLK}/50$ or $f_{CLK}/100$. To implement this example for the MF4-50 the clock frequency will have to be set to f_{CLK} = 50(1.184 kHz) = 59.2 kHz, or for the MF4-100, f_{CLK} = 100 (1.184 kHz) = 118.4 kHz.

2.2 CASCADING MF4s

When a steeper stopband attenuation rate is required, two MF4s can be cascaded (*Figure 7*) yielding an 8th order

2.0 Designing With The MF4 (Continued)

slope of 48 dB per octave. Because the MF4 is a Butterworth filter and therefore has no ripple in its passband when MF4s are cascaded, the resulting filter also has no ripple in its passband. Likewise the DC and passband gains will remain at 1V/V. The resulting response is shown in *Figure 9*.

In determining whether the cascaded MF4s will yield a filter that will meet a particular amplitude response specification, as above, equations 3 and 4 can be used, shown below.

$$n = \frac{\log[(10^{0.05A_{min}} - 1)/(10^{0.05A_{max}} - 1)]}{2 \log (f_s/f_c)} \quad (2)$$

$$\text{Attn}(f) = 10 \log [1 + (10^{0.05A_{max}} - 1)(f/f_c)^{2n}] \text{ dB} \quad (3)$$

where $n = 4$ (the order of each filter).

Equation 2 will determine whether the order of the filter is adequate ($n \leq 4$) while equation 3 can determine the actual stopband attenuation and cutoff frequency (f_c) necessary to obtain the desired frequency response. The design procedure would be identical to the one shown in section 2.0.

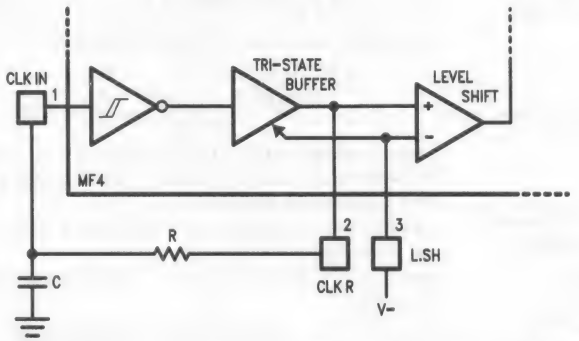
2.3 CHANGING CLOCK FREQUENCY INSTANTANEOUSLY

The MF4 will respond favorably to an instantaneous change in clock frequency. If the control signal in *Figure 9* is low the MF4-50 has a 100 kHz clock making $f_c = 2$ kHz; when this signal goes high the clock frequency changes to 50 kHz yielding $f_c = 1$ kHz. As the Figure illustrates, the output signal changes quickly and smoothly in response to a sudden change in clock frequency.

The step response of the MF4 in *Figure 10* is dependent on f_c. The MF4 responds as a classical fourth-order Butterworth low-pass filter.

2.4 ALIASING CONSIDERATIONS

Aliasing effects have to be considered when input signal frequencies exceed half the sampling rate. For the MF4 this equals half the clock frequency (f_{CLK}). When the input signal contains a component at a frequency higher than half the clock frequency $f_{CLK}/2$, as in *Figure 11a*, that component will be "reflected" about $f_{CLK}/2$ into the frequency range below $f_{CLK}/2$, as in *Figure 11b*. If this component is within the passband of the filter and of large enough amplitude it can cause problems. Therefore, if frequency components in the input signal exceed $f_{CLK}2$ they must be attenuated before being applied to the MF4 input. The necessary amount of attenuation will vary depending on system requirements. In critical applications the signal components above $f_{CLK}/2$ will have to be attenuated at least to the filter's residual noise level.



$$f = \frac{1}{RC \ln\left[\left(\frac{V_{CC} - V_{T-}}{V_{CC} - V_{T+}}\right)\left(\frac{V_{T+}}{V_{T-}}\right)\right]}$$

$$f \cong \frac{1}{1.69\,RC}$$

$(V_{CC} = 10V)$

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FIGURE 1. Schmitt Trigger R/C Oscillator

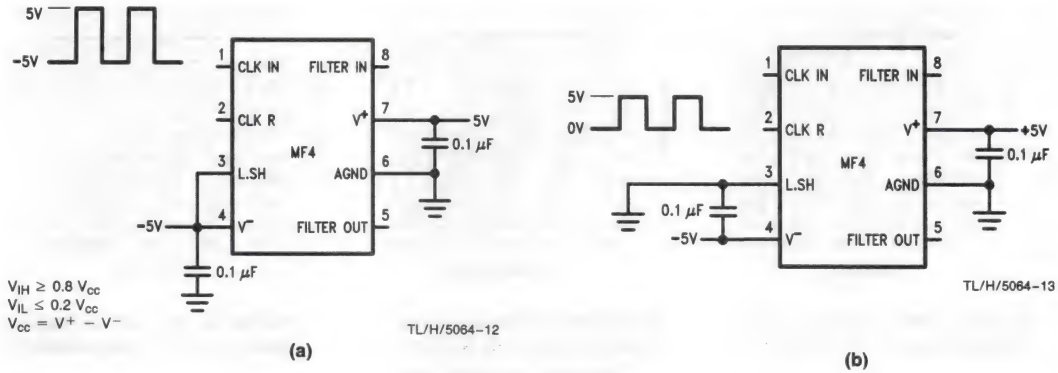


FIGURE 2. Split Supply Operation with CMOS Level Clock (a) and TTL Level Clock (b)

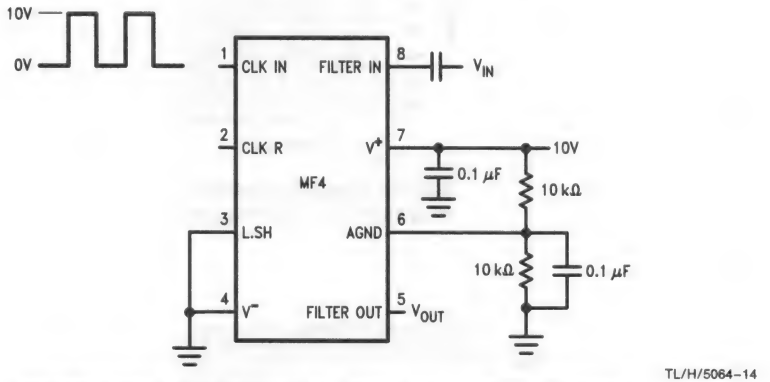
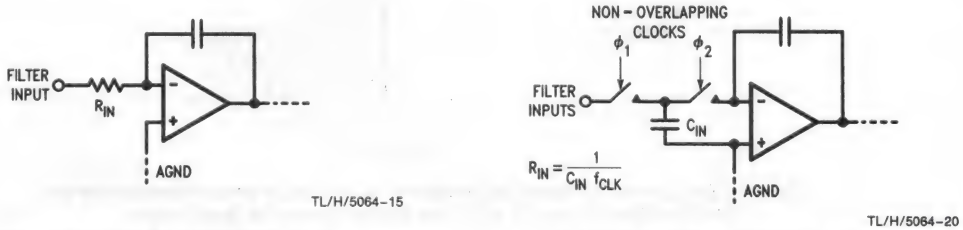


FIGURE 3. Single Supply Operation. ANGD Resistor Biased to $V^+/2$



a) Equivalent Circuit for MF4 Filter Input **b) Actual Circuit for MF4 Filter Input**

FIGURE 4. MF4 Filter Input

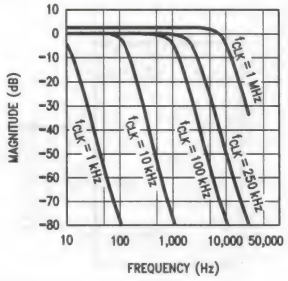


FIGURE 5a. MF4-100 Amplitude Response with ±5V Supplies

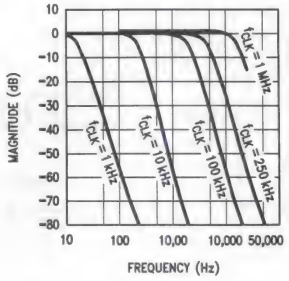


FIGURE 5b. MF4-50 Amplitude Response with ±5V Supplies

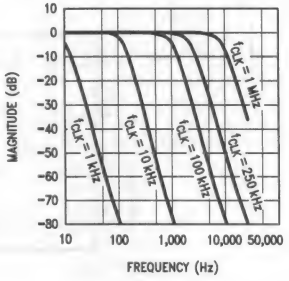


FIGURE 5c. MF4-100 Amplitude Response with ±2.5V Supplies

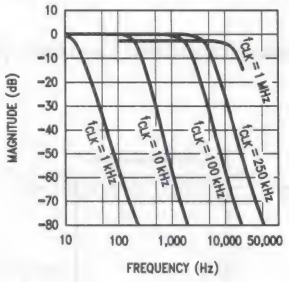


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FIGURE 5d. MF4-50 Amplitude Response with ±2.5V Supplies

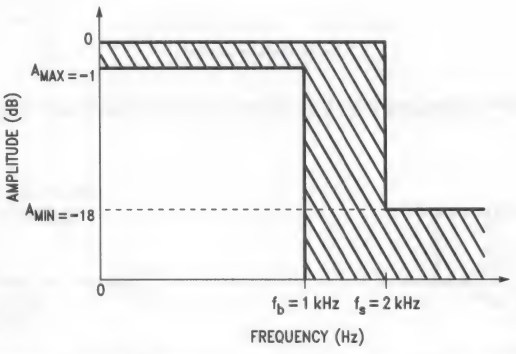

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FIGURE 6. Design Example Magnitude Response Specification where the Response of the Filter Design must fall within the shaded area of the specification

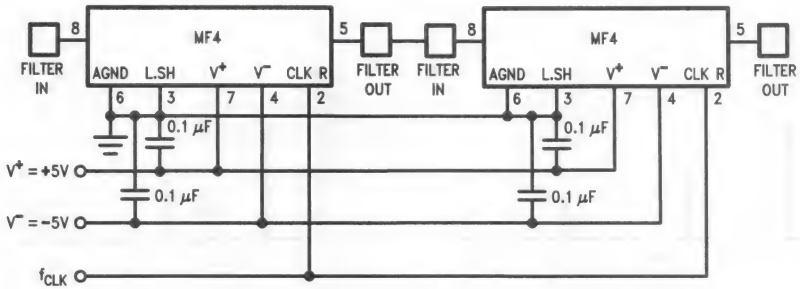

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FIGURE 7. Cascading Two MF4s

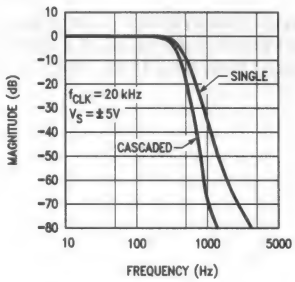

FIGURE 8a. One MF4-50 vs Two MF4-50s Cascaded

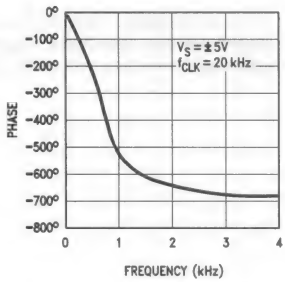

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FIGURE 8b. Phase Response of Two Cascaded MF4-50s

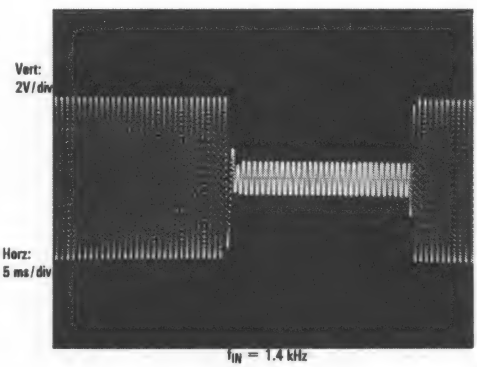

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FIGURE 9. MF4-50 Abrupt Clock Frequency Change

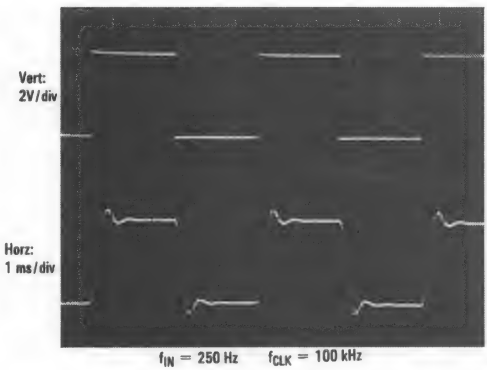

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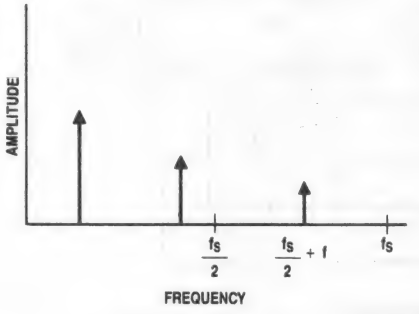
FIGURE 10. MF4-50 Input Step Response



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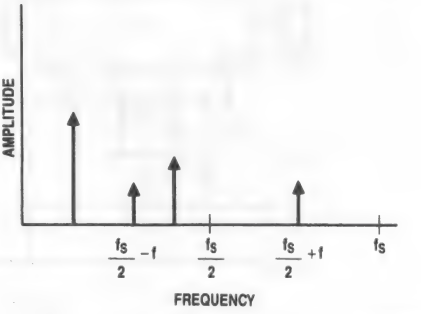
(a) input signal spectrum



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(b) Output signal spectrum. Note that the input signal at $f_c/2 + f$ causes an output signal to appear at $f_c/2 - f$.

FIGURE 11. The phenomenon of aliasing in sampled-data systems. An input signal whose frequency is greater than one-half the sampling frequency will cause an output to appear at a frequency lower than one-half the sampling frequency. In the MF4, $f_s = f_{CLK}$.

National Semiconductor

MF5 Universal Monolithic Switched Capacitor Filter

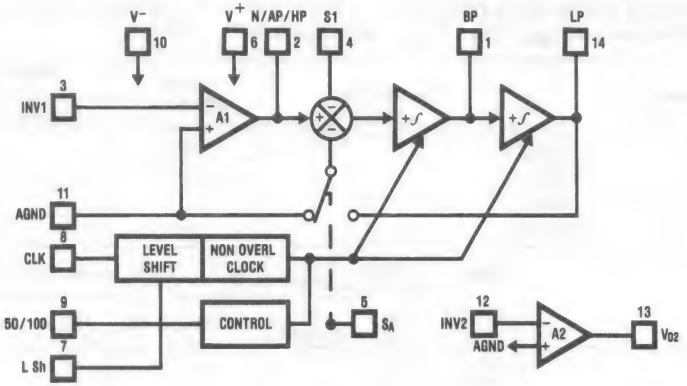
General Description

The MF5 consists of an extremely easy to use, general purpose CMOS active filter building block and an uncommitted op amp. The filter building block, together with an external clock and a few resistors, can produce various second order functions. The filter building block has 3 output pins. One of the output pins can be configured to perform highpass, allpass or notch functions and the remaining 2 output pins perform bandpass and lowpass functions. The center frequency of the filter can be directly dependent on the clock frequency or it can depend on both clock frequency and external resistor ratios. The uncommitted op amp can be used for cascading purposes, for obtaining additional allpass and notch functions, or for various other applications. Higher order filter functions can be obtained by cascading several MF5s or by using the MF5 in conjuction with the MF10 (dual switched capacitor filter building block). The MF5 is functionally compatible with the MF10. Any of the classical filter configurations (such as Butterworth, Bessel, Cauer and Chebyshev) can be formed.

Features

- Low cost
- 14-pin DIP or 14-pin Surface Mount (SO) wide-body package
- Easy to use
- Clock to center frequency ratio accuracy ±0.6%
- Filter cutoff frequency stability directly dependent on external clock quality
- Low sensitivity to external component variations
- Separate highpass (or notch or allpass), bandpass, lowpass outputs
- $f_o \times Q$ range up to 200 kHz
- Operation up to 30 kHz (typical)
- Additional uncommitted op-amp

Block and Connection Diagrams



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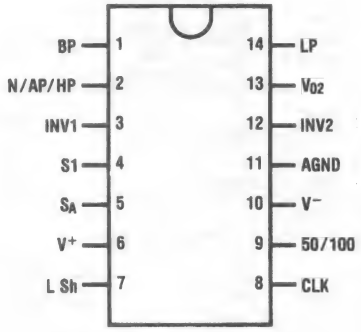
All Packages



Top View

TL/H/5066–2

Order Number MF5CN
See NS Package Number N14A

Order Number MF5CWM
See NS Package Number M14B

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage ($V^+ - V^-$)	14V
Power Dissipation $T_A = 25°C$ (note 1)	500 mW
Storage Temp.	150°C
Soldering Information:	
N Package: 10 sec.	260°C
SO Package: Vapor phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

Input Voltage (any pin)	$V^- \le V_{in} \le V^+$
Operating Temp. Range	$T_{MIN} \le T_A \le T_{MAX}$
MF5CN, MF5CWM	$0°C \le T_A \le 70°C$

Electrical Characteristics

$V^+ = 5V \pm 0.5\%$, $V^- = -5V \pm 0.5\%$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \le T_A \le T_{MAX}$.** For all other limits $T_A = 25°C$.

Parameter		Conditions	Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
Supply Voltage ($V^+ - V^-$)	Min				**8**	V
	Max				**14**	V
Maximum Supply Current		Clock applied to Pin 8 No Input Signal	4.5	6.0		mA
Clock Feedthrough	Filter Output		10			mV
	Op-amp Output		10			mV

Filter Electrical Characteristics

$V^+ = 5V \pm 0.5\%$, $V^- = -5V \pm 0.5\%$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \le T_A \le T_{MAX}$.** For all other limits $T_A = 25°C$.

Parameter		Conditions		Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
Center Frequency Range (f_o)	Max			30		**20**	kHz
	Min			0.1		**0.2**	Hz
Clock Frequency Range (f_{CLK})	Max			1.5		**1.0**	MHz
	Min			5.0		**10**	Hz
Clock to Center Frequency Ratio (f_{CLK}/f_o)		Ideal Q = 10 Mode 1	$V_{pin9} = +5V$ $F_{CLK} = 250$ kHz	50.11 ± 0.2%	50.11 ± 1.5%		
			$V_{pin9} = -5V$ $F_{CLK} = 500$ kHz	100.04 ± 0.2%	100.04 ± 1.5%		
f_{CLK}/f_o Temp. Coefficient		$V_{pin9} = +5V$ (50:1 CLK ratio)		±10			ppm/°C
		$V_{pin9} = -5V$ (100:1 CLK ratio)		±20			ppm/°C
Q Accuracy (Max) (Note 2)		Ideal Q = 10 Mode 1	$V_{pin9} = +5V$ $F_{CLK} = 250$ kHz		±6		%
			$V_{pin9} = -5V$ $F_{CLK} = 500$ kHz		±6		%
Q Temperature Coefficient		$V_{pin9} = +5V$ (50:1 CLK ratio)		−200			ppm/°C
		$V_{pin9} = -5V$ (100:1 CLK ratio)		−70			ppm/°C
DC Lowpass Gain Accuracy (Max)		Mode 1 R1 = R2 = 10 kΩ			±0.2		dB
DC Offset Voltage (Max) (Note 3)	V_{os1}			±5.0			mV
	V_{os2}	$V_{pin9} = +5V$ (50:1 CLK ratio)		−185			mV
	V_{os3}			+115			mV
	V_{os2}	$V_{pin9} = -5V$ (100:1 CLK ratio)		−310			mV
	V_{os3}			+240			mV

Filter Electrical Characteristics

$V^+ = 5V \pm 0.5\%$, $V^- = -5V \pm 0.5\%$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \leq T_A \leq T_{MAX}$.** For all other limits $T_A = 25°C$. (Continued)

Parameter		Conditions	Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
Output Swing (Min)	BP, LP pins	RL = 5 kΩ	±4.0	±3.8		V
	N/AP/HP pin	RL = 3.5 kΩ	±4.2	±3.8		V
Dynamic Range (Note 4)		$V_{pin9} = +5V$ (50:1 CLK ratio)	83			dB
		$V_{pin9} = -5V$ (100:1 CLK ratio)	80			dB
Maximum Output Short Circuit Current (Note 5)	Source		20			mA
	Sink		3.0			mA

OP-AMP Electrical Characteristics

$V^+ = +5V \pm 0.5\%$, $V^- = -5V \pm 0.5\%$ unless other noted. **Boldface limits apply over temperature, $T_{MIN} \leq T_A \leq T_{MAX}$.** For all other limits $T_A = 25°C$.

Parameter		Conditions	Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
Gain Bandwidth Product			2.5			MHz
Output Voltage Swing (Min)		RL = 3.5 kΩ	±4.2	±3.8		V
Slew Rate			7.0			V/μs
DC Open-Loop Gain			80			db
Input Offset Voltage (Max)			±5.0	±20		mV
Input Bias Current			10			pA
Maximum Output Short Circuit Current (Note 5)	Source		20			mA
	Sink		3.0			mA

Logic Input Characteristics

Boldface limits apply over temperature, $T_{MIN} \leq T_A \leq T_{MAX}$. All other limits $T_A = 25°C$.

Parameter		Conditions	Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
CMOS Clock Input	Min Logical "1" Input Voltage	$V^+ = +5V, V^- = -5V,$ $V_{L.Sh.} = 0V$		3.0		V
	Max Logical "0" Input Voltage			−3.0		V
	Min Logical "1" Input Voltage	$V^+ = +10V, V^- = 0V,$ $V_{L.Sh.} = +5V$		8.0		V
	Max Logical "0" Input Voltage			2.0		V
TTL Clock Input	Min Logical "1" Input Voltage	$V^+ = +5V, V^- = -5V,$ $V_{L.Sh.} = 0V$		2.0		V
	Max Logical "0" Input Voltage			0.8		V

Note 1: The typical junction-to-ambient thermal resistance (θ_{JA}) of the 14 pin N package is 160°C/W, and 82°C/W for the M package.

Note 2: The accuracy of the Q value is a function of the center frequency (f_o). This is illustrated in the curves under the heading "Typical Performance Characteristics".

Note 3: V_{os1}, V_{os2}, and V_{os3} refer to the internal offsets as discussed in the Application Information section 3.4.

Note 4: For ±5V supplies the dynamic range is referenced to 2.82V rms (4V peak) where the wideband noise over a 20 kHz bandwidth is typically 200 μV rms for the MF5 with a 50:1 CLK ratio and 280 μV rms for the MF5 with a 100:1 CLK ratio.

Note 5: The short circuit source current is measured by forcing the output that is being tested to its maximum positive voltage swing and then shorting that output to the negative supply. The short circuit sink current is measured by forcing the output that is being tested to its maximum negative voltage swing and then shorting that output to the positive supply. These are the worst case conditions.

Note 6: Typicals are at 25°C and represent most likely parametric norm.

Note 7: Guaranteed and 100% tested.

Note 8: Guaranteed, but not 100% tested. These limits are not used to calculate outgoing quality levels.

Pin Description

LP(14), BP(1), N/AP/HP(2): The second order lowpass, bandpass, and notch/allpass/highpass outputs. The LP and BP outputs can typically sink 1 mA and source 3 mA. The N/AP/HP output can typically sink 1.5 mA and source 3 mA. Each output typically swings to within 1V of each supply.

INV1(3): The inverting input of the summing op amp of the filter. This is a high impedance input, but the non-inverting input is internally tied to AGND, making INV1 behave like a summing junction (low impedance current input).

S1(4): S1 is a signal input pin used in the allpass filter configurations (see modes 4 and 5). The pin should be driven with a source impedance of less than 1 kΩ. If S1 is not driven with a signal it should be tied to AGND (mid-supply).

SA(5): This pin activates a switch that connects one of the inputs of the filter's second summer to either AGND (SA tied to V−) or to the lowpass (LP) output (SA tied to V+). This offers the flexibility needed for configuring the filter in its various modes of operation.

50/100(9): This pin is used to set the internal clock to center frequency ratio (f_{CLK}/f_o) of the filter. By tying the pin to V+ an f_{CLK}/f_o ratio of about 50:1 (typically 50.11 ± 0.2%) is obtained. Tying the 50/100 pin to either AGND or V− will set the f_{CLK}/f_o ratio to about 100:1 (typically 100.04 ± 0.2%).

AGND(11): This is the analog ground pin. This pin should be connected to the system ground for dual supply operation or biased to mid-supply for single supply operation. For a further discussion of mid-supply biasing techniques see the Applications Information (Section 3.2). For optimum filter performance a "clean" ground must be provided.

V+(6), V−(10): These are the positive and negative supply pins. The MF5 will operate over a total supply range of 8V to 14V. Decoupling the supply pins with 0.1 μF capacitors is highly recommended.

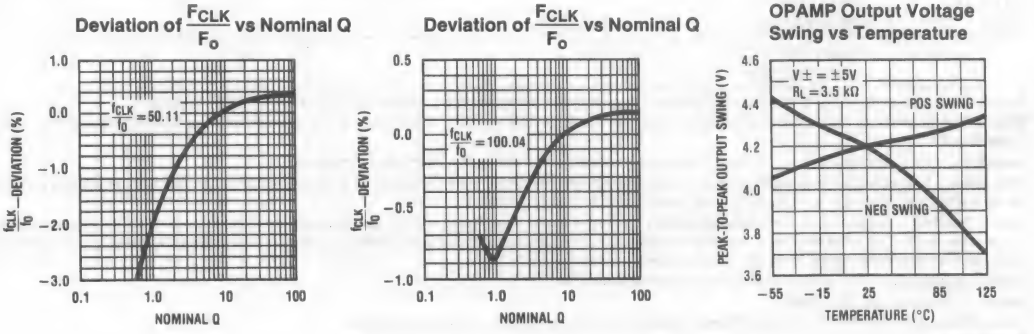
CLK(8): This is the clock input for the filter. CMOS or TTL logic level clocks can be accomodated by setting the L. Sh pin to the levels described in the L. Sh pin description. For optimum filter performance a 50% duty cycle clock is recommended for clock frequencies greater than 200 kHz. This gives each op amp the maximum amount of time to settle to a new sampled input.

L. Sh(7): This pin allows the MF5 to accommodate either CMOS or TTL logic level clocks. For dual supply operation (i.e., ±5V), a CMOS or TTL logic level clock can be accepted if the L. Sh pin is tied to mid-supply (AGND), which should be the system ground. For single supply operation the L. Sh pin should be tied to mid-supply (AGND) for a CMOS logic level clock. The mid-supply bias should be a very low impedance node. See Applications Information for biasing techniques. For a TTL logic level clock the L. Sh pin should be tied to V− which should be the system ground.

INV2(12): This is the inverting input of the uncommitted op amp. This is a very high impedance input, but the non-inverting input is internally tied to AGND, making INV2 behave like a summing junction (low-impedance current input).

Vo2(13): This is the output of the uncommitted op amp. It will typically sink 1.5 mA and source 3.0 mA. It will typically swing to within 1V of each supply.

Typical Performance Characteristics



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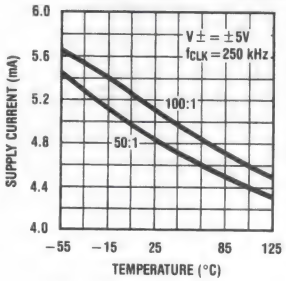
Typical Performance Characteristics (Continued)

Supply Current vs Temperature



TL/H/5066–4

1.0 Definitions of Terms

f_{CLK}: the frequency of the external clock signal applied to pin 8.

f_o: center frequency of the second order function complex pole pair. f_o is measured at the bandpass output of the MF5, and is the frequency of maximum bandpass gain. (*Figure 1*).

f_{notch}: the frequency of minimum (ideally zero) gain at the notch output.

f_z: the center frequency of the second order complex zero pair, if any. If f_z is different from f_o and if Q_z is high, it can be observed as the frequency of a notch at the allpass output. (*Figure 10*).

Q: "quality factor" of the 2nd order filter. Q is measured at the bandpass output of the MF5 and is equal to f_o divided by the −3dB bandwidth of the 2nd order bandpass filter (*Figure 1*). The value of Q determines the shape of the 2nd order filter responses as shown in *Figure 6*.

Q_z: the quality factor of the second order complex zero pair, if any. Q_z is related to the allpass characteristic, which is written:

$$H_{AP}(s) = \frac{H_{OAP}\left(s^2 - \frac{s\omega_o}{Q_z} + \omega_o^2\right)}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

where $Q_z = Q$ for an all-pass response.

H_{OBP}: the gain (in V/V) of the bandpass output at $f = f_o$.

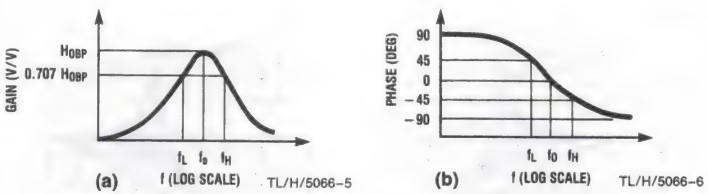
H_{OLP}: the gain (in V/V) of the lowpass output as $f \rightarrow 0$ Hz (*Figure 2*).

H_{OHP}: the gain (in V/V) of the highpass output as $f \rightarrow f_{clk}/2$ (*Figure 3*).

H_{ON}: the gain (in V/V) of the notch output as $f \rightarrow 0$ Hz and as $f \rightarrow f_{clk}/2$, when the notch filter has equal gain above and below the center frequency (*Figure 4*). When the low-frequency gain differs from the high-frequency gain, as in modes 2 and 3a (*Figures 11* and *8*), the two quantities below are used in place of H_{ON}.

H_{ON1}: the gain (in V/V) of the notch output as $f \rightarrow 0$ Hz.

H_{ON2}: the gain (in V/V) of the notch output as $f \rightarrow f_{clk}/2$.



$$H_{BP}(s) = \frac{H_{OBP}s}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

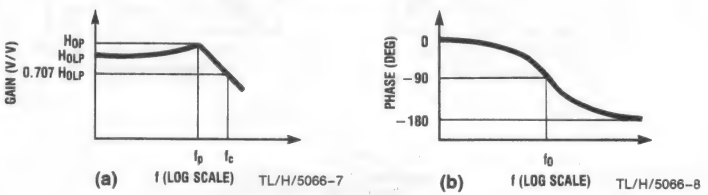
$$Q = \frac{f_o}{f_H - f_L};\quad f_o = \sqrt{f_L f_H}$$

$$f_L = f_o\left(\frac{-1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

$$f_H = f_o\left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

$$\omega_o = 2\pi f_o$$

FIGURE 1. 2nd-Order Bandpass Response



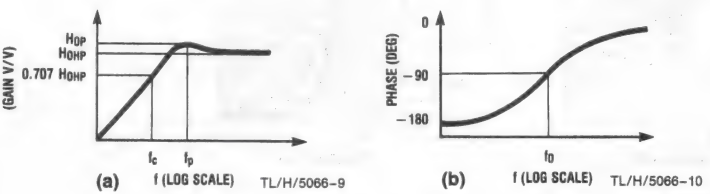
$$H_{LP}(s) = \frac{H_{OLP}\omega_o^2}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

$$f_c = f_o \times \sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}$$

$$f_p = f_o\sqrt{1 - \frac{1}{2Q^2}}$$

$$H_{OP} = H_{OLP} \times \frac{1}{\frac{1}{Q}\sqrt{1 - \frac{1}{4Q^2}}}$$

FIGURE 2. 2nd-Order Low-Pass Response



$$H_{HP}(s) = \frac{H_{OHP}s^2}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

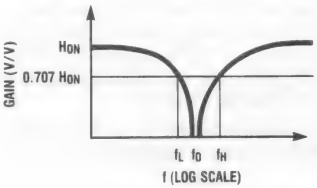
$$f_c = f_o \times \left[\sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}\right]^{-1}$$

$$f_p = f_o \times \left[\sqrt{1 - \frac{1}{2Q^2}}\right]^{-1}$$

$$H_{OP} = H_{OHP} \times \frac{1}{\frac{1}{Q}\sqrt{1 - \frac{1}{4Q^2}}}$$

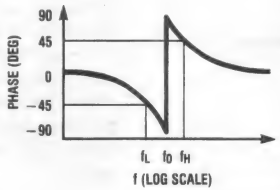
FIGURE 3. 2nd-Order High-Pass Response

1.0 Definition of Terms (Continued)



TL/H/5066–11

(a)



TL/H/5066–12

(b)

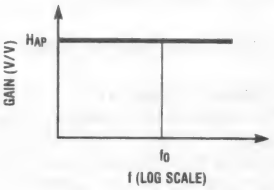
$$H_N(s) = \frac{H_{ON}(s^2 + \omega_o^2)}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

$$Q = \frac{f_o}{f_H - f_L}; \quad f_o = \sqrt{f_L f_H}$$

$$f_L = f_o\left(\frac{-1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

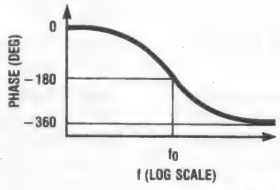
$$f_H = f_o\left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

FIGURE 4. 2nd-Order Notch Response



TL/H/5066–13

(a)



TL/H/5066–14

(b)

$$H_{AP}(s) = \frac{H_{OAP}\left(s^2 - \frac{s\omega_o}{Q} + \omega_o^2\right)}{s^2 + \frac{s\omega_o}{Q} + \omega_o^2}$$

FIGURE 5. 2nd-Order All-Pass Response

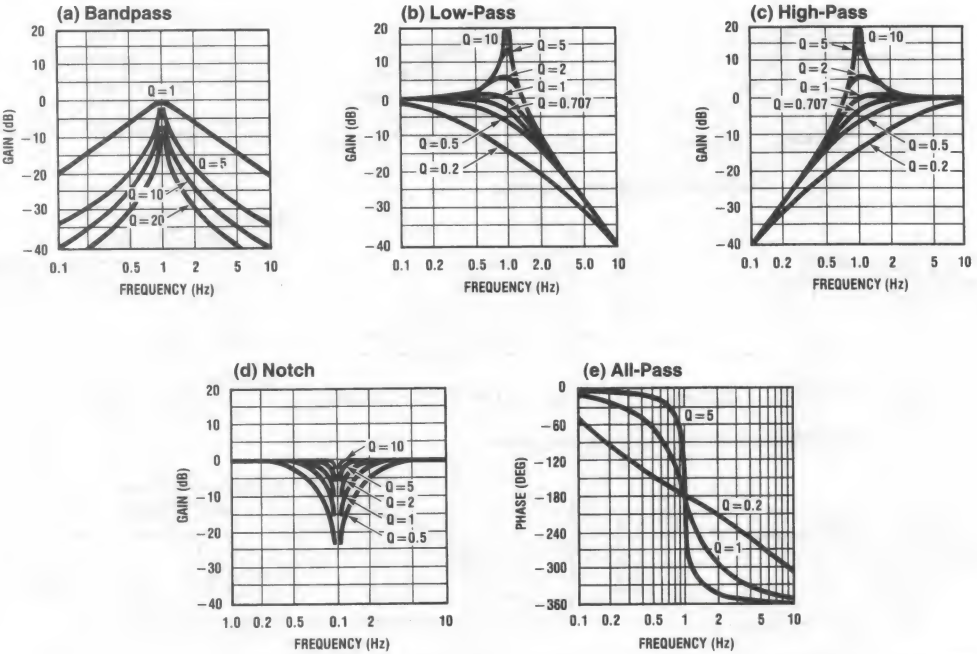


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FIGURE 6. Responses of various 2nd-order filters as a function of Q. Gains and center frequencies are normalized to unity.

2.0 Modes of Operation

The MF5 is a switched capacitor (sampled data) filter. To fully describe its transfer functions, a time domain approach is appropriate. Since this is cumbersome, and since the MF5 closely approximates continuous filters, the following discussion is based on the well known frequency domain. Each MF5 can produce a full 2nd order function. See Table 1 for a summary of the characteristics of the various modes.

MODE 1: Notch 1, Bandpass, Lowpass Outputs:

$f_{notch} = f_o$ **(See *Figure 7*)**

f_o = center frequency of the complex pole pair

$= \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

f_{notch} = center frequency of the imaginary zero pair = f_o.

H_{OLP} = Lowpass gain (as $f \rightarrow 0$) $= -\frac{R2}{R1}$

H_{OBP} = Bandpass gain (at $f = f_o$) $= -\frac{R3}{R1}$

H_{ON} = Notch output gain as $f \rightarrow 0$, $f \rightarrow f_{CLK}/2$ $= \frac{-R_2}{R_1}$

$Q = \frac{f_o}{BW} = \frac{R3}{R2}$

BW = the −3 dB bandwidth of the bandpass output.

Circuit dynamics:

$$H_{OLP} = \frac{H_{OBP}}{Q} \text{ or } H_{OBP} = H_{OLP} \times Q = H_{ON} \times Q.$$

$H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

MODE 1a: Non-Inverting BP, LP (See *Figure 8*)

$f_o = \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

$Q = \frac{R3}{R2}$

$H_{OLP} = -1$; $H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

$H_{OBP_1} = -\frac{R3}{R2}$

$H_{OBP_2} = 1$ (non-inverting)

Circuit dynamics: $H_{OBP_1} = Q$

Note: V_{IN} should be driven from a low impedance (<1 kΩ)

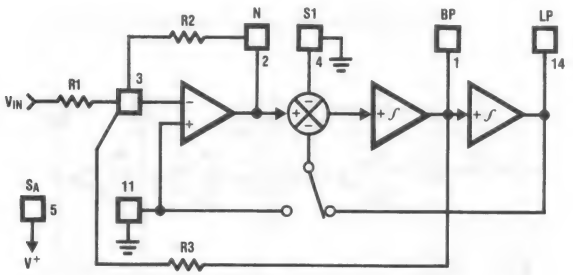

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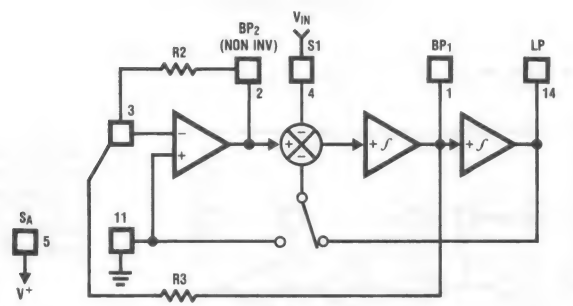
FIGURE 7. MODE 1



TL/H/5066–17

FIGURE 8. MODE 1a

2.0 Modes of Operation (Continued)

MODE 2: Notch 2, Bandpass, Lowpass: $f_{notch} < f_o$
(See *Figure 9*)

f_o = center frequency
$= \frac{f_{CLK}}{100}\sqrt{\frac{R2}{R4}+1}$ or $\frac{f_{CLK}}{50}\sqrt{\frac{R2}{R4}+1}$

$f_{notch} = \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

Q = quality factor of the complex pole pair
$= \frac{\sqrt{R2/R4+1}}{R2/R3}$

H_{OLP} = Lowpass output gain (as $f \rightarrow 0$)
$= -\frac{R2/R1}{R2/R4+1}$

H_{OBP} = Bandpass output gain (at $f = f_o$) $= -R3/R1$

H_{ON_1} = Notch output gain (as $f \rightarrow 0$)
$= -\frac{R2/R1}{R2/R4+1}$

H_{ON_2} = Notch output gain $\left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right) = -R2/R1$

Filter dynamics: $H_{OBP} = Q\sqrt{H_{OLP}\,H_{ON_2}} = Q\sqrt{H_{ON_1}\,H_{ON_2}}$

MODE 3: Highpass, Bandpass, Lowpass Outputs
(See *Figure 10*)

$f_o = \frac{f_{CLK}}{100} \times \sqrt{\frac{R2}{R4}}$ or $\frac{f_{CLK}}{50} \times \sqrt{\frac{R2}{R4}}$

Q = quality factor of the complex pole pair
$= \sqrt{\frac{R2}{R4}} \times \frac{R3}{R2}$

H_{OHP} = Highpass gain $\left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right) = -\frac{R2}{R1}$

H_{OBP} = Bandpass gain (at $f = f_o$) $= -\frac{R3}{R1}$

H_{OLP} = Lowpass gain (as $f \rightarrow 0$) $= -\frac{R4}{R1}$

Circuit dynamics: $\frac{R2}{R4} = \frac{H_{OHP}}{H_{OLP}}$; $H_{OBP} = \sqrt{H_{OHP} \times H_{OLP}} \times Q$

$H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)
$H_{OHP(peak)} \cong Q \times H_{OHP}$ (for high Q's)

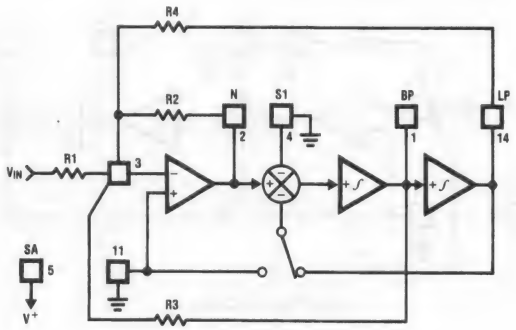


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FIGURE 9. MODE 2

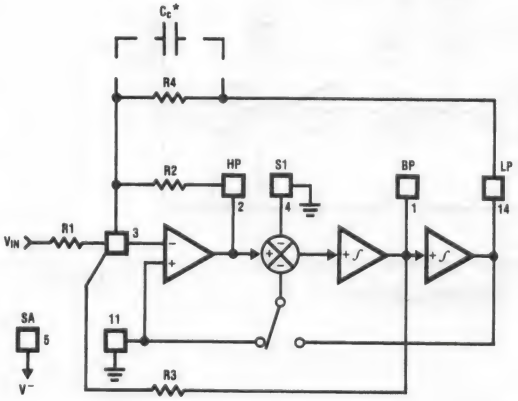

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FIGURE 10. MODE 3

*In Mode 3, the feedback loop is closed around the input summing amplifier; the finite GBW product of this op amp causes a slight Q enhancement. If this is a problem, connect a small capacitor (10 pF–100 pF) across R4 to provide some phase lead.

2.0 Modes of Operation (Continued)

MODE 3a: HP, BP, LP and Notch with External Op amp (See *Figure 11*)

$$f_o = \frac{f_{CLK}}{100} \times \sqrt{\frac{R2}{R4}} \text{ or } \frac{f_{CLK}}{50} \times \sqrt{\frac{R2}{R4}}$$

$$Q = \sqrt{\frac{R2}{R4}} \times \frac{R3}{R2}$$

$$H_{OHP} = -\frac{R2}{R1}$$

$$H_{OBP} = -\frac{R3}{R1}$$

$$H_{OLP} = -\frac{R4}{R1}$$

f_n = notch frequency $= \frac{f_{CLK}}{100}\sqrt{\frac{R_h}{R_l}}$ or $\frac{f_{CLK}}{50}\sqrt{\frac{R_h}{R_l}}$

H_{on} = gain of notch at $f = f_o = \left\| Q\left(\frac{R_g}{R_l} H_{OLP} - \frac{R_g}{R_h} H_{OHP}\right)\right\|$

H_{n1} = gain of notch (as $f \rightarrow 0$) $= \frac{R_g}{R_l} \times H_{OLP}$

H_{n2} = gain of notch $\left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right) = -\frac{R_g}{R_h} \times H_{OHP}$

MODE 4: Allpass, Bandpass, Lowpass Outputs (See *Figure 12*)

f_o = center frequency

$= \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$;

f_z^* = center frequency of the complex zero pair $\cong f_o$

$$Q = \frac{f_o}{BW} = \frac{R3}{R2};$$

Q_z = quality factor of complex zero pair $= \frac{R3}{R1}$

For AP output make R1 = R2

H^*_{OAP} = Allpass gain $\left(\text{at } 0 < f < \frac{f_{CLK}}{2}\right) = -\frac{R2}{R1} = -1$

H_{OLP} = Lowpass gain (as $f \rightarrow 0$)

$= -\left(\frac{R2}{R1} + 1\right) = -2$

H_{OBP} = Bandpass gain (at $f = f_o$)

$= -\frac{R3}{R2}\left(1 + \frac{R2}{R1}\right) = -2\left(\frac{R3}{R2}\right)$

Circuit dynamics: $H_{OBP} = (H_{OLP}) \times Q = (H_{OAP} + 1)\, Q$

*Due to the sampled data nature of the filter, a slight mismatch of f_z and f_o occurs causing a 0.4 dB peaking around f_o of the allpass filter amplitude response (which theoretically should be a straight line). If this is unacceptable, Mode 5 is recommended.

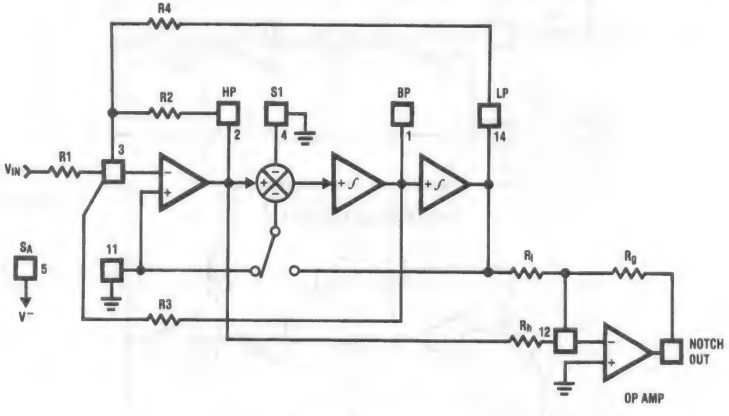

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FIGURE 11. MODE 3a

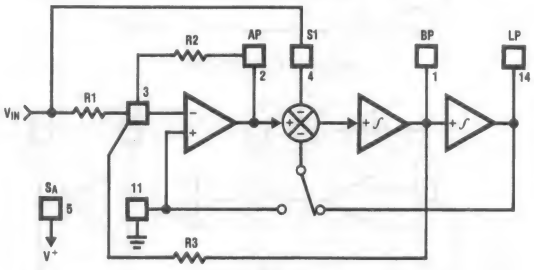

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FIGURE 12. MODE 4

2.0 Modes of Operation (Continued)

MODE 5: Numerator Complex Zeros, BP, LP (See *Figure 13*)

$$f_o = \sqrt{1+\frac{R2}{R4}} \times \frac{f_{CLK}}{100} \text{ or } \sqrt{1+\frac{R2}{R4}} \times \frac{f_{CLK}}{50}$$

$$f_z = \sqrt{1-\frac{R1}{R4}} \times \frac{f_{CLK}}{100} \text{ or } \sqrt{1-\frac{R1}{R4}} \times \frac{f_{CLK}}{50}$$

$$Q = \sqrt{1+R2/R4} \times \frac{R3}{R2}$$

$$Q_z = \sqrt{1-R1/R4} \times \frac{R3}{R1}$$

$$H_{0z1} = \text{gain at C.Z. output (as } f \rightarrow 0 \text{ Hz)} = \frac{-R2\,(R4-R1)}{R1\,(R4+R2)}$$

$$H_{0z2} = \text{gain at C.Z. output } \left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right) = \frac{-R2}{R1}$$

$$H_{OBP} = -\left(=\frac{R2}{R1}+1\right) \times \frac{R3}{R2}$$

$$H_{OLP} = -\left(\frac{R2+R1}{R2+R4}\right) \times \frac{R4}{R1}$$

MODE 6a: Single Pole, HP, LP Filter (See *Figure 14*)

f_c = cutoff frequency of LP or HP output

$$= \frac{R2}{R3}\frac{f_{CLK}}{100} \text{ or } \frac{R2}{R3}\frac{f_{CLK}}{50}$$

$$H_{OLP} = -\frac{R3}{R1}$$

$$H_{OHP} = -\frac{R2}{R1}$$

MODE 6b: Single Pole LP Filter (Inverting and Non-Inverting) (See *Figure 15*)

f_c = cutoff frequency of LP outputs

$$\cong \frac{R2}{R3}\frac{f_{CLK}}{100} \text{ or } \frac{R2}{R3}\frac{f_{CLK}}{50}$$

H_{OLP_1} = 1 (non-inverting)

$$H_{OLP_2} = -\frac{R3}{R2}$$

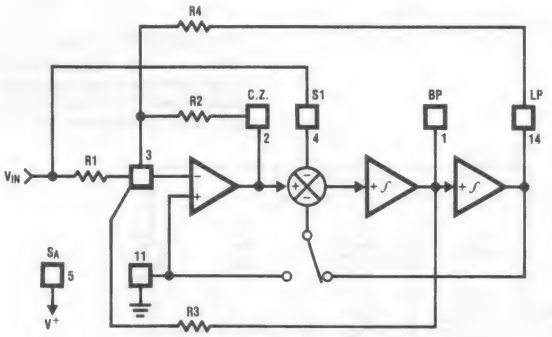

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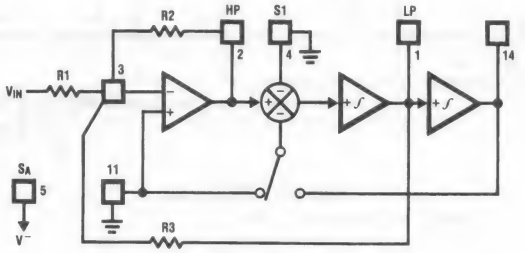
FIGURE 13. MODE 5



TL/H/5066–23

FIGURE 14. MODE 6a

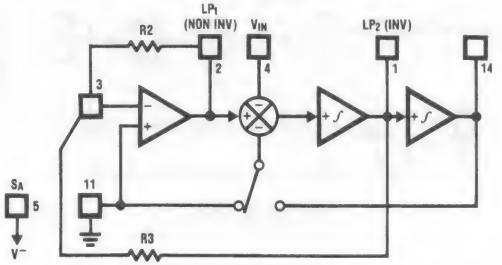


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FIGURE 15. MODE 6b

2.0 Modes of Operation (Continued)

TABLE I. Summary of Modes. Realizable filter types (e.g. low-pass) denoted by asterisks. Unless otherwise noted, gains of various filter outputs are inverting and adjustable by resistor ratios.

Mode	BP	LP	HP	N	AP	Number of resistors	Adjustable f_{CLK}/f_o	Notes
1	*	*		*		3	No	
1a	(2) $H_{OBP1}=-Q$ $H_{OBP2}=+1$	$H_{OLP}=+1$				2	No	May need input buffer. Poor dynamics for high Q.
2	*	*		*		3	Yes (above $f_{CLK}/50$ or $f_{CLK}/100$)	
3	*	*	*			4	Yes	Universal State-Variable Filter. Best general-purpose mode.
3a	*	*	*	*		7	Yes	As above, but also includes resistor-tuneable notch.
4	*	*			*	3	No	Gives Allpass response with $H_{OAP}=-1$ and $H_{OLP}=-2$.
5	*	*			*	4		Gives flatter allpass response than above if $R_1=R_2=0.02R_4$.
6a		*	*			3		Single pole.
6b		(2) $H_{OLP}=+1$ $H_{OLP2}=\frac{-R3}{R2}$				2		Single pole

3.0 Applications Information

The MF5 is a general-purpose second-order state variable filter whose center frequency is proportional to the frequency of the square wave applied to the clock input (f_{CLK}). By connecting pin 9 to the appropriate DC voltage, the filter center frequency f_o can be made equal to either $f_{CLK}/100$ or $f_{CLK}/50$. f_o can be very accurately set (within ±0.6%) by using a crystal clock oscillator, or can be easily varied over a wide frequency range by adjusting the clock frequency. If desired, the f_{CLK}/f_o ratio can be altered by external resistors as in *Figures 9, 10, 11, 13, 14*, and *15*. The filter Q and gain are determined by external resistors.

All of the five second-order filter types can be built using the MF5. These are illustrated in *Figures 1* through *5* along with their transfer functions and some related equations. *Figure 6* shows the effect of Q on the shapes of these curves. When filter orders greater than two are desired, two or more MF5s can be cascaded. The MF5 also includes an uncommitted CMOS operational amplifier for additional signal processing applications.

3.1 DESIGN EXAMPLE

An example will help illustrate the MF5 design procedure. For the example, we will design a 2nd order Butterworth low-pass filter with a cutoff frequency of 200 Hz, and a passband gain of −2. The circuit will operate from a ±5V power supply, and the clock amplitude will be ±5v (CMOS) levels).

From the specifications, the filter parameters are: $f_o=200$ Hz, $H_{OLP}=-2$, and, for Butterworth response, $Q=0.707$.

In section 2.0 are several modes of operation for the MF5, each having different characteristics. Some allow adjustment of f_{CLK}/f_o, others produce different combinations of filter types, some are inverting while others are non-inverting, etc. These characteristics are summarized in Table I. To keep the example simple, we will use mode 1, which has notch, bandpass, and lowpass outputs, and inverts the signal polarity. Three external resistors determine the filter's Q and gain. From the equations accompanying *Figure 7*, $Q=R_3/R_2$ and the passband gain $H_{OLP}=-R_2/R_1$. Since the input signal is driving a summing junction through R_1, the input impedance will be equal to R_1. Start by choosing a value for R_1. 10k is convenient and gives a reasonable input impedance. For $H_{OLP}=-2$, we have:

$R_2=-R_1H_{OLP}=10k\times 2=20k$.

For $Q=0.707$ we have:

$R_3=R_2Q=20k\times 0.707=14.14k$. Use 15k.

For operation on ±5V supplies, V^+ is connected to +5V, V^- to −5V, and AGND to ground. The power supplies should be "clean" (regulated supplies are preferred) and 0.1 μF bypass capacitors are recommended.

3.0 Applications Information (Continued)

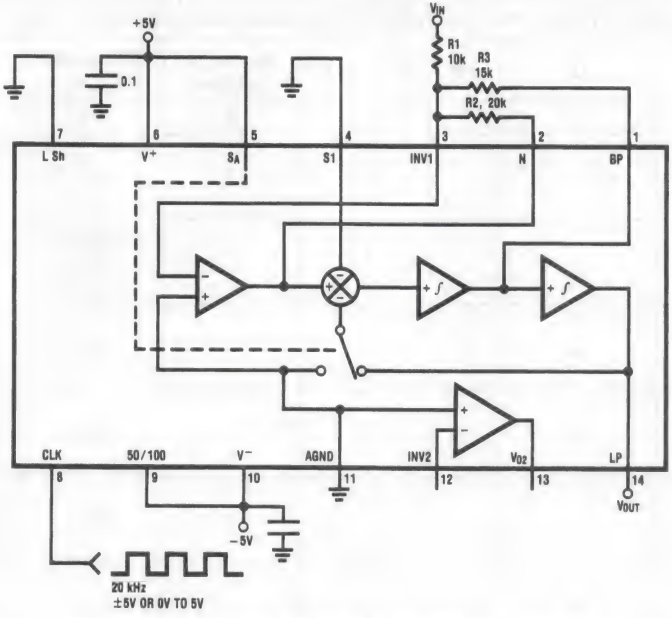

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FIGURE 16. 2nd-Order Butterworth Low-Pass Filter of Design Example. For $\frac{f_{CLK}}{f_0} = 50$, Connect Pin 9 to +5V, and Change Clock Frequency to 10 kHz.

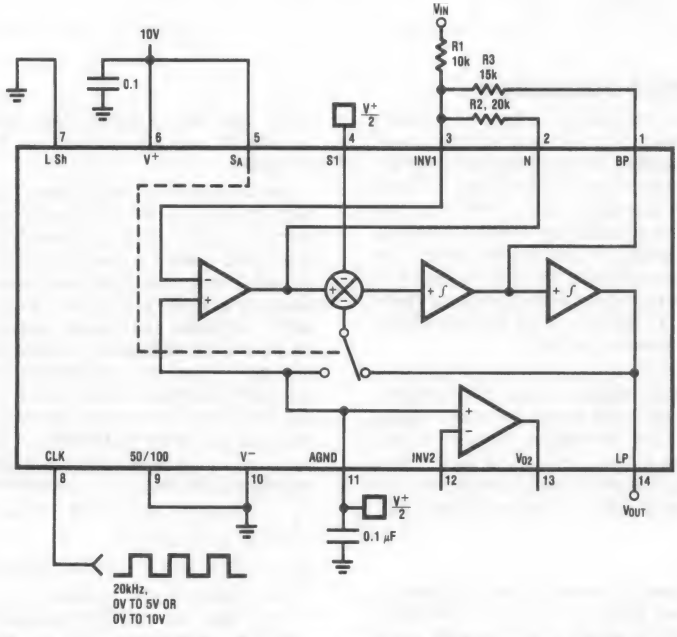

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FIGURE 17. Butterworth Low-Pass Circuit of Example, but Designed for Single-Supply Operation

3.0 Applications Information (Continued)

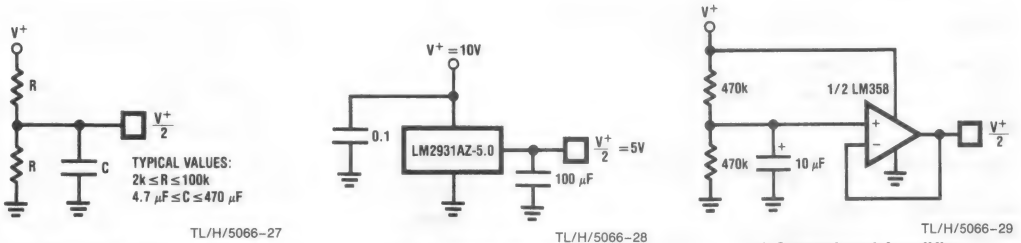

FIGURE 18. Three Ways of Generating $\frac{V^+}{2}$ for Single-supply Operation

For a cutoff frequency of 200 Hz, the external clock can be either 10 kHz with pin 9 connected to V^+ (50:1) or 20 kHz with pin 9 tied to A_{GND} or V^- (100:1). The voltage on the Logic Level Shift pin (7) determines the logic threshold for the clock input. The threshold is approximately 2V higher than the voltage applied to pin 7. Therefore, when pin 7 is grounded, the clock logic threshold will be 2V, making it compatible with 0–5 volt TTL logic levels and ±5 volt CMOS levels. Pin 7 should be connected to a clean, low-impedance (less than 1000Ω) voltage source.

The complete circuit of the design example is shown for a 100:1 clock ratio in *Figure 16*.

3.2 SINGLE SUPPLY OPERATION

The MF5 can also operate with a single-ended power supply. *Figure 17* shows the example filter with a single-ended power supply. V^+ is again connected to the positive power supply (8 to 14 volts), and V^- is connected to ground. The A_{GND} pin must be tied to $V^+/2$ for single supply operation. This half-supply point should be very "clean", as any noise appearing on it will be treated as an input to the filter. It can be derived from the supply voltage with a pair of resistors and a bypass capacitor (*Figure 18a*), or a low-impedance half-supply voltage can be made using a three-terminal voltage regulator or an operational amplifier (*Figures 18b* and *18c*). The passive resistor divider with a bypass capacitor is sufficient for many applications, provided that the time constant is long enough to reject any power supply noise. It is also important that the half-supply reference present a low impedance to the clock frequency, so at very low clock frequencies the regulator or op-amp approaches may be preferable because they will require smaller capacitors to filter the clock frequency. The main power supply voltage should be clean (preferably regulated) and bypassed with 0.1μF.

3.3 DYNAMIC CONSIDERATIONS

The maximum signal handling capability of the MF5, like that of any active filter, is limited by the power supply voltages used. The amplifiers in the MF5 are able to swing to within about 1 volt of the supplies, so the input signals must be kept small enough that none of the outputs will exceed these limits. If the MF5 is operating on ±5 volts, for example, the outputs will clip at about $8V_{p-p}$. The maximum input voltage multiplied by the filter gain should therefore be less than $8V_{p-p}$.

Note that if the filter has high Q, the gain at the lowpass or highpass outputs will be much greater than the nominal filter gain (*Figure 6*). As an example, a lowpass filter with a Q of 10 will have a 20 dB peak in its amplitude response at f_o. If the nominal gain of the filter H_{OLP} is equal to 1, the gain at f_o will be 10. The maximum input signal at f_o must therefore be less than 800 mV_{p-p} when the circuit is operated on ±5 volt supplies.

Also note that one output can have a reasonable small voltage on it while another is saturated. This is most likely for a circuit such as the notch in Mode 1 (*Figure 7*). The notch output will be very small at f_o, so it might appear safe to apply a large signal to the input. However, the bandpass will have its maximum gain at f_o and can clip if overdriven. If one output clips, the performance at the other outputs will be degraded, so avoid overdriving any filter section, even ones whose outputs are not being directly used. Accompanying *Figures 7* through *15* are equations labeled "circuit dynamics", which relate the Q and the gains at the various outputs. These should be consulted to determine peak circuit gains and maximum allowable signals for a given application.

3.4 OFFSET VOLTAGE

The MF5's switched capacitor integrators have a higher equivalent input offset voltage than would be found in a typical continuous-time active filter integrator. *Figure 19* shows an equivalent circuit of the MF5 from which the output dc offsets can be calculated. Typical values for these offsets are:

V_{os1} = opamp offset = ±5mV

V_{os2} = −185mV @ 50:1 −310mV @ 100:1

V_{os3} = +115mV @ 50:1 +240mV @ 100:1

The dc offset at the BP output is equal to the input offset of the lowpass integrator (V_{os3}). The offsets at the other outputs depend on the mode of operation and the resistor ratios, as described in the following expressions.

3.0 Applications Information (Continued)

Mode 1 and Mode 4

$$V_{OS(N)} = V_{OS1}\left(\frac{1}{Q} + 1 + \left\|H_{OLP}\right\|\right) - \frac{V_{OS3}}{Q}$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N)} - V_{OS2}$$

Mode 1a

$$V_{OS(N.INV.BP)} = \left(1 + \frac{1}{Q}\right) V_{OS1} - \frac{V_{OS3}}{Q}$$

$$V_{OS(INV.BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N.INV.BP)} - V_{OS2}$$

Mode 2 and Mode 5

$$V_{OS(N)} = \left(\frac{R2}{Rp} + 1\right) V_{OS1} \times \frac{1}{1 + R2/R4} + V_{OS2}\frac{1}{1 + R4/R2} - \frac{V_{OS3}}{Q\sqrt{1 + R2/R4}};$$

$$R_p = R1//R2//R4$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N)} - V_{OS2}$$

Mode 3

$$V_{OS(HP)} = V_{OS2}$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = -\frac{R4}{R2}\left(\frac{R2}{R3}V_{OS3} + V_{OS2}\right) + -\frac{R4}{R2}\left(1 + \frac{R2}{R_p}\right) V_{OS1};\ R_p = R1//R3//R4$$

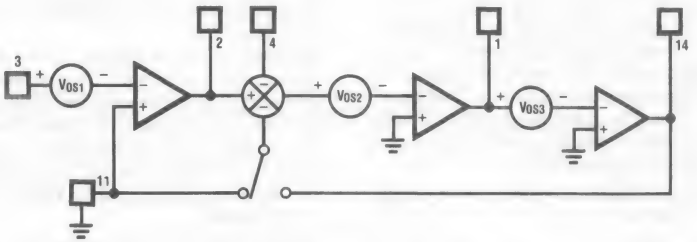

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FIGURE 19. Block Diagram Showing MF5 Offset Voltage Sources

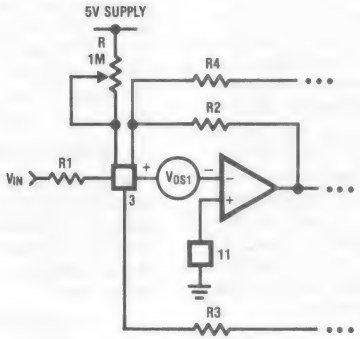

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FIGURE 20. Method for Trimming V_{OS}, See Text, Section 3.4

3.0 Applications Information (Continued)

For most applications, the outputs are AC coupled and DC offsets are not bothersome unless large signals are applied to the filter input. However, larger offset voltages will cause clipping to occur at lower ac signal levels, and clipping at any of the outputs will cause gain nonlinearities and will change f_o and Q. When operating in Mode 3, offsets can become excessively large if R_2 and R_4 are used to make f_{CLK}/f_o significantly higher than the nominal value, especially if Q is also high. An extreme example is a bandpass filter having unity gain, a Q of 20, and $f_{CLK}/f_o = 250$ with pin 9 tied to V^- (100:1 nominal). R_4/R_2 will therefore be equal to 6.25 and the offset voltage at the lowpass output will be about +1.9V. Where necessary, the offset voltage can be adjusted by using the circuit of *Figure 20*. This allows adjustment of V_{OS1}, which will have varying effects on the different outputs as described in the above equations. Some outputs cannot be adjusted this way in some modes, however ($V_{OS(BP)}$ in modes 1a and 3, for example).

3.5 SAMPLED DATA SYSTEM CONSIDERATIONS

The MF5 is a sampled data filter, and as such, differs in many ways from conventional continuous-time filters. An important characteristic of sampled-data systems is their effect on signals at frequencies greater than one-half the sampling frequency. (The MF5's sampling frequency is the same as its clock frequency). If a signal with a frequency greater than one-half the sampling frequency is applied to the input of a sampled data system, it will be "reflected" to a frequency less than one-half the sampling frequency. Thus, an input signal whose frequency is $f_s/2 + 100$ Hz will cause the system to respond as though the input frequency was $f_s/2$ - 100 Hz. This phenomenon is known as "aliasing", and can be reduced or eliminated by limiting the input signal spectrum to less than $f_s/2$. This may in some cases require the use of a bandwidth-limiting filter ahead of the MF5 to limit the input spectrum. However, since the clock frequency is much higher than the center frequency, this will often not be necessary.

Another characteristic of sampled-data circuits is that the output signal changes amplitude once every sampling period, resulting in "steps" in the output voltage which occur at the clock rate. (*Figure 21*) If necessary, these can be "smoothed" with a simple R-C low-pass filter at the MF5 output.

The ratio of f_{CLK} to f_c (normally either 50:1 or 100:1) will also affect performance. A ratio of 100:1 will reduce any aliasing problems and is usually recommended for wideband input signals. In noise sensitive applications, however, a ratio of 50:1 may be better as it will result in 3 dB lower output noise. The 50:1 ratio also results in lower DC offset voltages, as discussed in 3.4.

The accuracy of the f_{CLK}/f_o ratio is dependent on the value of Q. This is illustrated in the curves under the heading "Typical Performance Characteristics". As Q is changed, the true value of the ratio changes as well. Unless the Q is low, the error in f_{CLK}/f_o will be small. If the error is too large for a specific application, use a mode that allows adjustment of the ratio with external resistors.

It should also be noted that the product of Q and f_o should be limited to 300 kHz when $f_o < 5$ kHz, and to 200 kHz for $f_o > 5$ kHz.

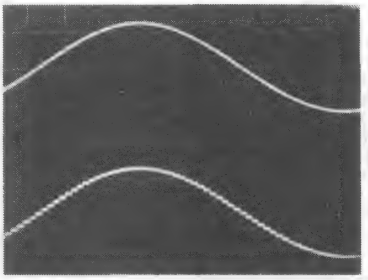

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FIGURE 21. The Sampled-Data Output Waveform

National Semiconductor

MF6 6th Order Switched Capacitor Butterworth Lowpass Filter

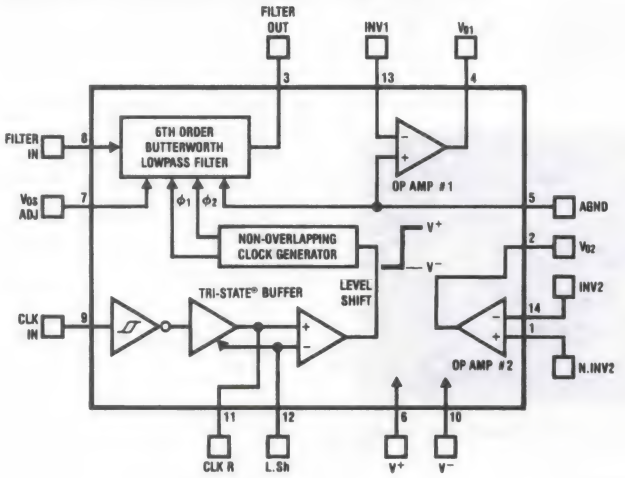
General Description

The MF6 is a versatile easy to use, precision 6th order Butterworth lowpass active filter. Switched capacitor techniques eliminate external component requirements and allow a clock tunable cutoff frequency. The ratio of the clock frequency to the lowpass cutoff frequency is internally set to 50 to 1 (MF6-50) or 100 to 1 (MF6-100). A Schmitt trigger clock input stage allows two clocking options, either self-clocking (via an external resistor and capacitor) for stand-alone applications, or an external TTL or CMOS logic compatible clock can be used for tighter cutoff frequency control. The maximally flat passband frequency response together with a DC gain of 1 V/V allows cascading MF6 sections for higher order filtering. In addition to the filter, two independent CMOS op amps are included on the die and are useful for any general signal conditioning applications.

Features

- No external components
- 14-pin DIP or 14-pin wide-body S.O. package
- Cutoff frequency accuracy of ±0.3% typical
- Cutoff frequency range of 0.1 Hz to 20 kHz
- Two uncommitted op amps available
- 5V to 14V total supply voltage
- Cutoff frequency set by external or internal clock

Block and Connection Diagrams



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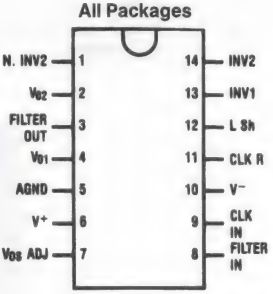
All Packages



TL/H/5065–2

Top View

Order Number MF6CWM-50 or MF6CWM-100
See NS Package Number M14B

Order Number MF6CN-50 or MF6CN-100
See NS Package Number N14A

Order Number MF6CJ-50 or MF6CJ-100
See NS Package Number J14A

Absolute Maximum Ratings (Note 11)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	14V
Voltage at Any Pin	$V^- - 0.2V$, $V^+ + 0.2V$
Input Current at Any Pin (Note 13)	5 mA
Package Input Current (Note 13)	20 mA
Power Dissipation (Note 14)	500 mW
Storage Temperature	−65°C to +150°C
ESD Susceptibility (Note 12)	800V
Soldering Information	
N Package (10 sec.)	260°C
J Package (10 sec.)	300°C
SO Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (Appendix D) for other methods of soldering surface mount devices.

Operating Ratings (Note 11)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
MF6CN-50, MF6CN-100	$0°C \le T_A \le +70°C$
MF6CWM-50, MF6CWM-100	$0°C \le T_A \le +70°C$
MF6CJ-50, MF6CJ-100	$-40°C \le T_A \le +85°C$
Supply Voltage ($V_S = V^+ - V^-$)	5V to 14V

Filter Electrical Characteristics

The following specifications apply for $f_{CLK} \le 250$ kHz (see Note 3) unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Parameter	Conditions	MF6CWM-50, MF6CWM-100, MF6CN-50, MF6CN-100			MF6CJ-50, MF6CJ-100			Units
		Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
$V^+ = +5V$, $V^- = -5V$								
f_c, Cutoff Frequency Range (Note 1) MF6-50 Min				**0.1**			**0.1**	Hz
MF6-50 Max				**20k**			**20k**	
MF6-100 Min				**0.1**			**0.1**	
MF6-100 Max				**10k**			**10k**	
Total Supply Current	f_{CLK} = 250 kHz	4.0	6.0	**8.5**	4.0	**8.5**		mA
Maximum Clock Feedthrough Filter Output		30			30			mV (peak-to-peak)
Op Amp 1 Out		25			25			
Op Amp 2 Out		20			20			
H_o, DC Gain	$R_{source} \le 2$ kΩ	0.0	±0.30	**±0.30**	0.0	**±0.30**		dB
f_{CLK}/f_c, Clock to Cutoff Frequency Ratio MF6-50		49.27 ± 0.3%	49.27 ± 1%	**49.27 ± 1%**	49.27 ± 0.3%	**49.27 ± 1%**		
MF6-100		98.97 ± 0.3%	98.97 ± 1%	**98.97 ± 1%**	98.97 ± 0.3%	**98.97 ± 1%**		
DC Offset Voltage MF6-50		−200			−200			mV
MF6-100		−400			−400			
Minimum Output Voltage Swing	R_L = 10 kΩ	+4.0	+3.5	**+3.5**	+4.0	**+3.5**		V
		−4.1	−3.8	**−3.5**	−4.1	**−3.5**		
Maximum Output Short Circuit Current (Note 6) Source		50	60	**80**	50	**80**		mA
Sink		1.5	2.0	**3.0**	1.5	**3.0**		
Dynamic Range (Note 2) MF6-50		83			83			dB
MF6-100		81			81			
Additional Magnitude Response Test Points (Note 4) MF6-50	f_{CLK} = 250 kHz							
	f = 6000 Hz	−9.47	−9.47 ± 0.5	**−9.47 ± 0.65**	−9.47	**−9.47 ± 0.65**		dB
	f = 4500 Hz	−0.92	−0.92 ± 0.2	**−0.92 ± 0.3**	−0.92	**−0.92 ± 0.3**		
MF6-100	f_{CLK} = 250 kHz							
	f = 3000 Hz	−9.48	−9.48 ± 0.5	**−9.48 ± 0.65**	−9.48	**−9.48 ± 0.65**		dB
	f = 2250 Hz	−0.97	−0.97 ± 0.2	**−0.97 ± 0.3**	−0.97	**−0.97 ± 0.3**		

Filter Electrical Characteristics (Continued) The following specifications apply for $f_{CLK} \leq 250$ kHz (see Note 3) unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Parameter	Conditions	MF6CWM-50, MF6CWM-100 MF6CN-50, MF6CN-100			MF6CJ-50, MF6CJ-100			Units
		Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
$V^+ = +5V, V^- = -5V$ (Continued)								
Attenuation Rate MF6-50	$f_{CLK} = 250$ kHz $f_1 = 6000$ Hz $f_2 = 8000$ Hz		−36	**−36**		**−36**		dB/ octave
MF6-100	$f_{CLK} = 250$ kHz $f_1 = 3000$ Hz $f_2 = 4000$ Hz		−36	**−36**		**−36**		dB/ octave
$V^+ = +2.5V, V^- = -2.5V$								
f_c, Cutoff Frequency Range (Note 1) MF6-50 Min				**0.1**			**0.1**	Hz
Max				**10k**			**10k**	
MF6-100 Min				**0.1**			**0.1**	
Max				**5k**			**5k**	
Total Supply Current	$f_{CLK} = 250$ kHz	2.5	4.0	**4.0**	2.5	**4.0**		mA
Maximum Clock Feedthrough Filter Output		20			20			mV (peak-to-peak)
Op Amp 1 Out		15			15			
Op Amp 2 Out		10			10			
H_o, DC Gain	$R_{source} \leq 2$ kΩ	0.0	±0.30	**±0.30**	0.0	**±0.30**		dB
f_{CLK}/f_c, Clock to Cutoff Frequency Ratio MF6-50		49.10 ± 0.3%	49.10 ± 2%	**49.10 ± 3%**	49.10 ± 0.3%	**49.10 ± 3%**		
MF6-100		98.65 ± 0.3%	98.65 ± 2%	**98.65 ± 2.25%**	98.65 ± 0.3%	**98.65 ± 2.25%**		
DC Offset Voltage MF6-50		−200			−200			mV
MF6-100		−400			−400			
Minimum Output Voltage Swing	$R_L = 10$ kΩ	+1.5	+1.0	**+1.0**	+1.5	**+1.0**		V
		−2.2	−1.7	**−1.5**	−2.2	**−1.5**		
Maximum Output Short Circuit Current (Note 6) Source		28	40	**50**	28	**50**		mA
Sink		0.5	1.0	**1.5**	0.5	**1.5**		
Dynamic Range (Note 2)		77			77			dB
Additional Magnitude Response Test Points (Note 4) MF6-50	$f_{CLK} = 250$ kHz							
	$f = 6000$ Hz	−9.54	−9.54 ± 0.5	**−9.54 ± 0.65**	−9.54	**−9.54 ± 0.65**		dB
	$f = 4500$ Hz	−0.96	−0.96 ± 0.2	**−0.96 ± 0.3**	−0.96	**−0.96 ± 0.3**		
MF6-100	$f_{CLK} = 250$ kHz							
	$f = 3000$ Hz	−9.67	−9.67 ± 0.5	**−9.67 ± 0.65**	−9.67	**−9.67 ± 0.65**		dB
	$f = 2250$ Hz	−1.01	−1.01 ± 0.2	**−1.01 ± 0.3**	−1.01	**−1.01 ± 0.3**		
Attenuation Rate MF6-50	$f_{CLK} = 250$ kHz $f_1 = 6000$ Hz $f_2 = 8000$ Hz		−36	**−36**		**−36**		dB/ octave
MF6-100	$f_{CLK} = 250$ kHz $f_1 = 3000$ Hz $f_2 = 4000$ Hz		−36	**−36**		**−36**		dB/ octave

Op Amp Electrical Characteristics

Boldface limits apply for T_{MIN} to T_{MAX}; all other limits $T_A = T_J = 25°C$.

Parameter	Conditions	MF6CN-50, MF6CN-100, MF6CWM-50, MF6CWM-100			MF6CJ-50, MF6CJ-100			Units
		Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
$V^+ = +5V, V^- = -5V$								
Input Offset Voltage		±8.0	±20	**±20**	±8.0	**±20**		mV
Input Bias Current		10			10			pA
CMRR (Op Amp #2 Only)	$V_{CM1} = 1.8V$, $V_{CM2} = -2.2V$	60	55		60	55		dB
Output Voltage Swing	$R_L = 10\ k\Omega$	+4.0 −4.5	+3.8 −4.0	**+3.6** **−4.0**	+4.0 −4.5	**+3.6** **−4.0**		V
Maximum Output Short Source Circuit Current (Note 6) Sink		54 2.0	65 4.0	**80** **6.0**	54 2.0	**80** **6.0**		mA
Slew Rate		7.0			7.0			V/μs
DC Open Loop Gain		72			72			dB
Gain Bandwidth Product		1.2			1.2			MHz
$V^+ = +2.5V, V^- = -2.5V$								
Input Offset Voltage		±8.0	±20	**±20**	±8.0	**±20**		mV
Input Bias Current		10			10			pA
CMRR (Op-Amp #2 Only)	$V_{CM1} = +0.5V$, $V_{CM2} = -0.9V$	60	55		60	55		dB
Output Voltage Swing	$R_L = 10\ k\Omega$	+1.5 −2.2	+1.3 −1.7	**+1.1** **−1.7**	+1.5 −2.2	**+1.1** **−1.7**		V
Maximum Output Short Source Circuit Current (Note 6) Sink		24 1.0	35 2.0	**50** **4.0**	24 1.0	**50** **4.0**		mA
Slew Rate		6.0			6.0			V/μs
DC Open Loop Gain		67			67			dB
Gain Bandwidth Product		1.2			1.2			MHz

Logic Input-Output Electrical Characteristics

The following specifications apply for $V^- = 0V$ (see Note 5) unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Parameter		Conditions		MF6CN-50, MF6CN-100 MF6CWM-50, MF6CWM-100			MF6CJ-50, MF6CJ-100			Units
				Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
TTL CLOCK INPUT, CLK R PIN (Note 7)										
Maximum V_{IL}, Logical "0" Input Voltage					0.8	**0.8**		**0.8**		V
Minimum V_{IH}, Logical "1" Input Voltage					2.0	**2.0**		**2.0**		V
Maximum Leakage Current at CLK R Pin		L Sh Pin at Mid- Supply			2.0	**2.0**		**2.0**		μA
SCHMITT TRIGGER										
V_{T+}, Positive Going Threshold Voltage	Min Max	$V^+ = 10V$		7.0	6.1 8.9	**6.1** **8.9**	7.0	**6.1** **8.9**		V
	Min Max	$V^+ = 5V$		3.5	3.1 4.4	**3.1** **4.4**	3.5	**3.1** **4.4**		V
V_{T-}, Negative Going Threshold Voltage	Min Max	$V^+ = 10V$		3.0	1.3 3.8	**1.3** **3.8**	3.0	**1.3** **3.8**		V
	Min Max	$V^+ = 5V$		1.5	0.6 1.9	**0.6** **1.9**	1.5	**0.6** **1.9**		V
Hysteresis ($V_{T+} - V_{T-}$)	Min Max	$V^+ = 10V$		4.0	2.3 7.6	**2.3** **7.6**	4.0	**2.3** **7.6**		V
	Min Max	$V^+ = 5V$		2.0	1.2 3.8	**1.2** **3.8**	2.0	**1.2** **3.8**		V
Minimum Logical "1" Output Voltage (Pin 11)		$I_o = -10\mu A$	$V^+ = 10V$ $V^+ = 5V$		9.0 4.5	**9.0** **4.5**		**9.0** **4.5**		V
Maximum Logical "0" Output Voltage (Pin 11)		$I_o = 10\mu A$	$V^+ = 10V$ $V^+ = 5V$		1.0 0.5	**1.0** **0.5**		**1.0** **0.5**		V
Minimum Output Source Current (Pin 11)		CLK R Tied to Ground	$V^+ = 10V$ $V^+ = 5V$	6.0 1.5	3.0 0.75	**3.0** **0.75**	6.0 1.5	**3.0** **0.75**		mA
Maximum Output Sink Current (Pin 11)		CLK R Tied to V^+	$V^+ = 10V$ $V^+ = 5V$	5.0 1.3	2.5 0.65	**2.5** **0.65**	5.0 1.3	**2.5** **0.65**		mA

Note 1: The cutoff frequency of the filter is defined as the frequency where the magnitude response is 3.01 dB less than the DC gain of the filter.

Note 2: For ±5V supplies the dynamic range is referenced to 2.82 Vrms (4V peak) where the wideband noise over a 20 kHz bandwidth is typically 200 μVrms for the MF6-50 and 250 μVrms for the MF6-100. For ±2.5V supplies the dynamic range is referenced to 1.06 Vrms (1.5V peak) where the wideband noise over a 20 kHz bandwidth is typically 140 μVrms for both the MF6-50 and the MF6-100.

Note 3: The specifications for the MF6 have been given for a clock frequency (f_{CLK}) of 250 kHz and less. Above this clock frequency the cutoff frequency begins to deviate from the specified error band of ±1.0% but the filter still maintains its magnitude characteristics. See Application Hints, Section 1.5.

Note 4: Besides checking the cutoff frequency (f_c) and the stopband attenuation at 2 f_c, two additional frequencies are used to check the magnitude response of the filter. The magnitudes are referenced to a DC gain of 0.0 dB.

Note 5: For simplicity all the logic levels have been referenced to $V^- = 0V$ and will scale accordingly for ±5V and ±2.5V supplies (except for the TTL input logic levels).

Note 6: The short circuit source current is measured by forcing the output that is being tested to its maximum positive voltage swing and then shorting that output to the negative supply. The short circuit sink current is measured by forcing the output that is being tested to its maximum negative voltage swing and then shorting that output to the positive supply. These are the worst-case conditions.

Note 7: The MF6 is operating with symmetrical split supplies and L.Sh is tied to ground.

Note 8: Typicals are at 25°C and represent most likely parametric norm.

Note 9: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level.

Note 10: Design limits are guaranteed, but not 100% tested. These limits are not used to calculate outgoing quality levels.

Note 11: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified conditions.

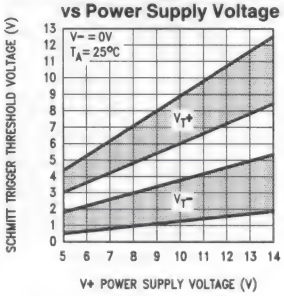
Note 12: Human body model, 100 pF discharged through a 1.5k Ω resistor.

Note 13: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 5 mA or less. The 20 mA package input current limits the number of pins that can exceed the power supply boundaries with a 5 mA current limit to four.

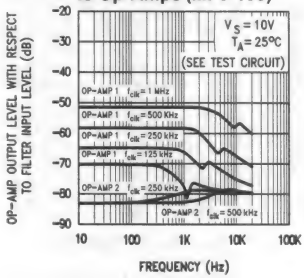
Note 14: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 125°C$, and the typical junction-to-ambient thermal resistance of the MF6CN when board mounted is 67°C/W. For the MF6CJ this number decreases to 62°C/W. For MF6CWM, $\theta_{JA} = 78°C/W$.

Typical Performance Characteristics

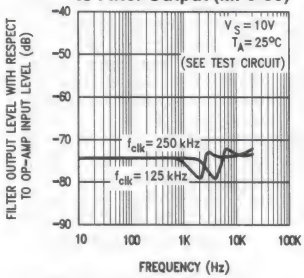
Schmitt Trigger Threshold Voltage vs Power Supply Voltage



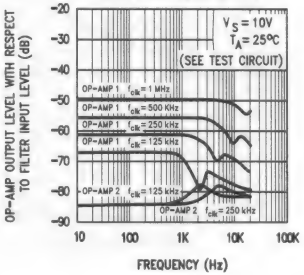
Crosstalk from Filter to Op-Amps (MF6-100)


VS = 10V
TA = 25°C
(SEE TEST CIRCUIT)
OP-AMP 1 fclk = 1 MHz
OP-AMP 1 fclk = 500 KHz
OP-AMP 1 fclk = 250 kHz
OP-AMP 1 fclk = 125 kHz
OP-AMP 2 fclk = 250 kHz
OP-AMP 2 fclk = 500 kHz
OP-AMP OUTPUT LEVEL WITH RESPECT TO FILTER INPUT LEVEL (dB)
FREQUENCY (Hz)

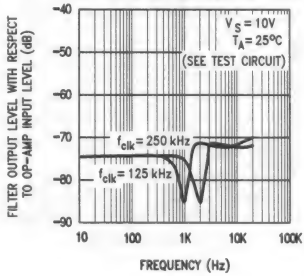
Crosstalk from Either Op-Amp to Filter Output (MF6-50)


VS = 10V
TA = 25°C
(SEE TEST CIRCUIT)
fclk = 250 kHz
fclk = 125 kHz
FILTER OUTPUT LEVEL WITH RESPECT TO OP-AMP INPUT LEVEL (dB)
FREQUENCY (Hz)

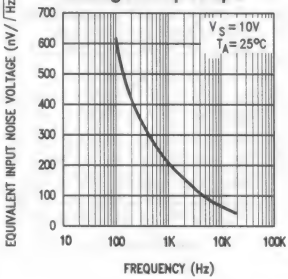
Crosstalk from Filter to Op-Amps (MF6-50)


VS = 10V
TA = 25°C
(SEE TEST CIRCUIT)
OP-AMP 1 fclk = 1 MHz
OP-AMP 1 fclk = 500 KHz
OP-AMP 1 fclk = 250 kHz
OP-AMP 1 fclk = 125 kHz
OP-AMP 2 fclk = 125 kHz
OP-AMP 2 fclk = 250 kHz
OP-AMP OUTPUT LEVEL WITH RESPECT TO FILTER INPUT LEVEL (dB)
FREQUENCY (Hz)

Crosstalk from Either Op-Amp to Filter Output (MF6-100)


VS = 10V
TA = 25°C
(SEE TEST CIRCUIT)
fclk = 250 kHz
fclk = 125 kHz
FILTER OUTPUT LEVEL WITH RESPECT TO OP-AMP INPUT LEVEL (dB)
FREQUENCY (Hz)

Equivalent Input Noise Voltage of Op-Amps



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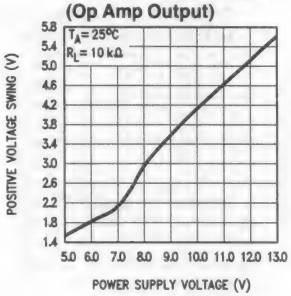
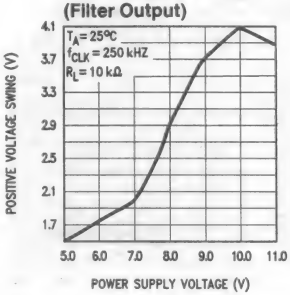
Typical Performance Characteristics (Continued)

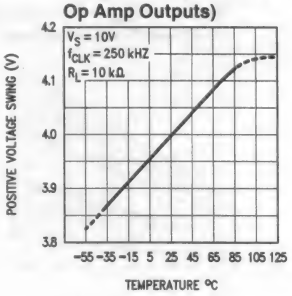
Positive Voltage Swing vs Power Supply Voltage (Op Amp Output)



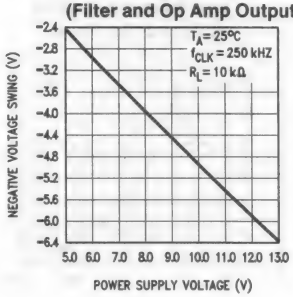
Positive Voltage Swing vs Power Supply Voltage (Filter Output)



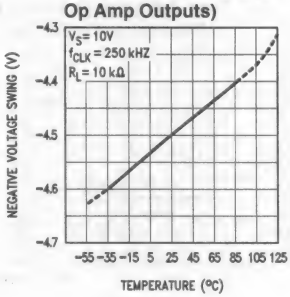
Positive Voltage Swing vs Temperature (Filter and Op Amp Outputs)



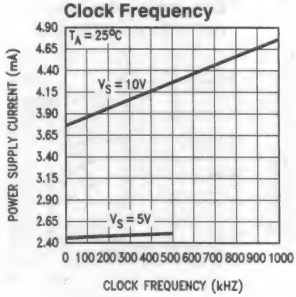
Negative Voltage Swing vs Power Supply Voltage (Filter and Op Amp Outputs)


T_A = 25°C
f_{CLK} = 250 kHZ
R_L = 10 kΩ
NEGATIVE VOLTAGE SWING (V)
−2.4 −2.8 −3.2 −3.6 −4.0 −4.4 −4.8 −5.2 −5.6 −6.0 −6.4
5.0 6.0 7.0 8.0 9.0 10.0 11.0 12.0 13.0
POWER SUPPLY VOLTAGE (V)

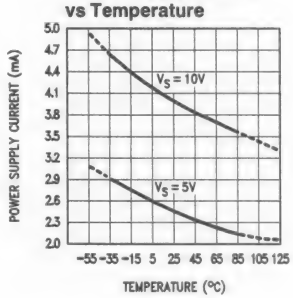
Negative Voltage Swing vs Temperature (Filter and Op Amp Outputs)



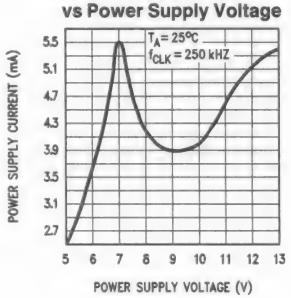
Power Supply Current vs Clock Frequency



Power Supply Current vs Temperature



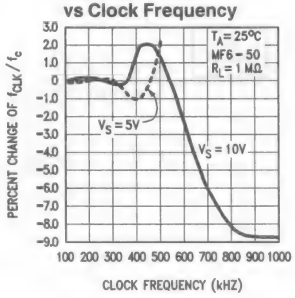
Power Supply Current vs Power Supply Voltage



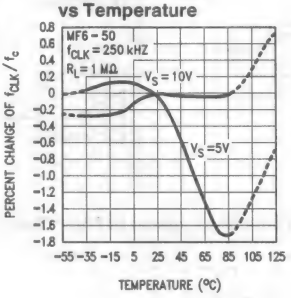
TL/H/5065–35

Typical Performance Characteristics (Continued)

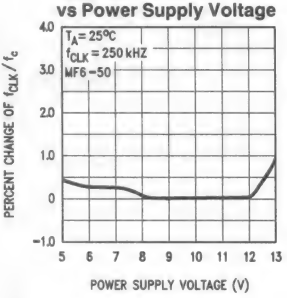
f_{CLK}/f_c Deviation vs Clock Frequency



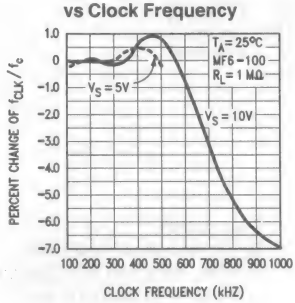
f_{CLK}/f_c Deviation vs Temperature



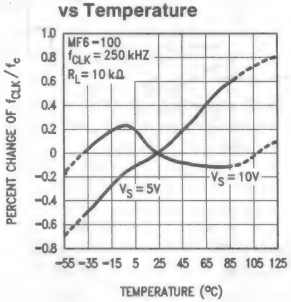
f_{CLK}/f_c Deviation vs Power Supply Voltage



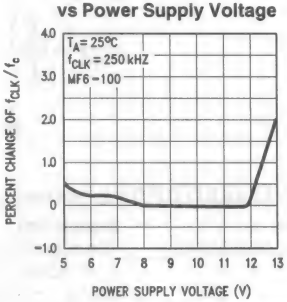
f_{CLK}/f_c Deviation vs Clock Frequency



f_{CLK}/f_c Deviation vs Temperature

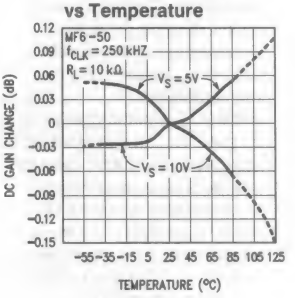


f_{CLK}/f_c Deviation vs Power Supply Voltage

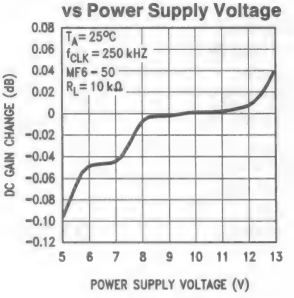


TL/H/5065–36

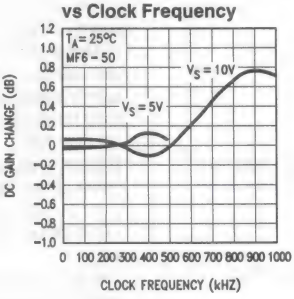
DC Gain Deviation vs Temperature



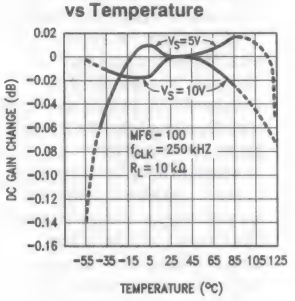
DC Gain Deviation vs Power Supply Voltage



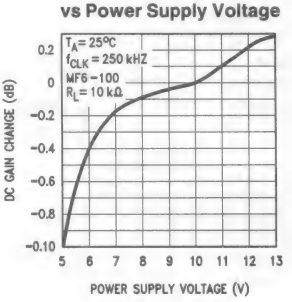
DC Gain Deviation vs Clock Frequency



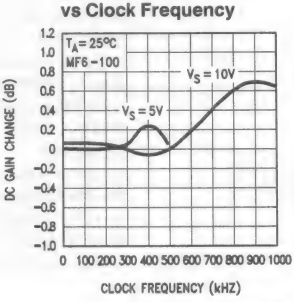
DC Gain Deviation vs Temperature



DC Gain Deviation vs Power Supply Voltage



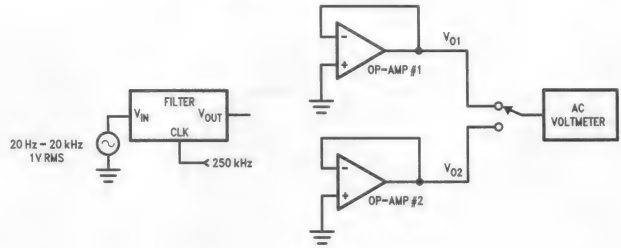
DC Gain Deviation vs Clock Frequency



TL/H/5065–39

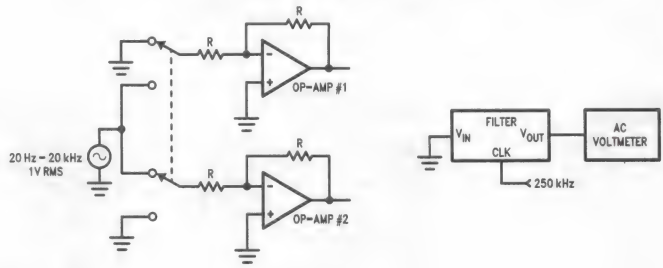
Crosstalk Test Circuits

From Filter to Opamps



TL/H/5065–10

From Either Opamp to Filter Output



TL/H/5065–11

Pin Descriptions (Pin Numbers)

Pin	Description
FILTER OUT (3)	The output of the lowpass filter. It will typically sink 0.9 mA and source 3 mA and swing to within 1V of each supply rail.
FILTER IN (8)	The input to the lowpass filter. To minimize gain errors the source impedance that drives this input should be less than 2k (see section 1.4). For single supply operation the input signal must be biased to mid-supply or AC coupled.
V_{OS}ADJ (7)	This pin is used to adjust the DC offset of the filter output; if not used it must be tied to the AGND potential. (See section 1.3)
AGND (5)	The analog ground pin. This pin sets the DC bias level for the filter section and the non-inverting input of Op-Amp #1 and must be tied to the system ground for split supply operation or to mid-supply for single supply operation (see section 1.2). When tied to mid-supply this pin should be well bypassed.
V_{O1} (4), INV1 (13)	V_{O1} is the output and INV1 is the inverting input of Op-Amp #1. The non-inverting input of this Op-Amp is internally connected to the AGND pin.
V_{O2} (2), INV2 (14), NINV2 (1)	V_{O2} is the output, INV2 is the inverting input, and NINV2 is the non-inverting input of Op-Amp #2.
V^+ (6), V^- (10)	The positive and negative supply pins. The total power supply range is 5V to 14V. Decoupling these pins with 0.1 μF capacitors is highly recommended.
CLK IN (9)	A CMOS Schmitt-trigger input to be used with an external CMOS logic level clock. Also used for self-clocking Schmitt-trigger oscillator (see section 1.1).
CLK R (11)	A TTL logic level clock input when in split supply operation ($\pm$2.5V to $\pm$7V) and L. Sh tied to system ground. This pin becomes a low impedance output when L. Sh is tied to V^-. Also used in conjunction with the CLK IN pin for a self clocking Schmitt-trigger oscillator (see section 1.1).
L. Sh (12)	Level shift pin, selects the logic threshold levels for the desired clock. When tied to V^- it enables an internal tri-state buffer stage between the Schmitt trigger and the internal clock level shift stage thus enabling the CLK IN Schmitt-trigger input and making the CLK R pin a low impedance output.

Pin Descriptions (Pin Numbers) (Continued)

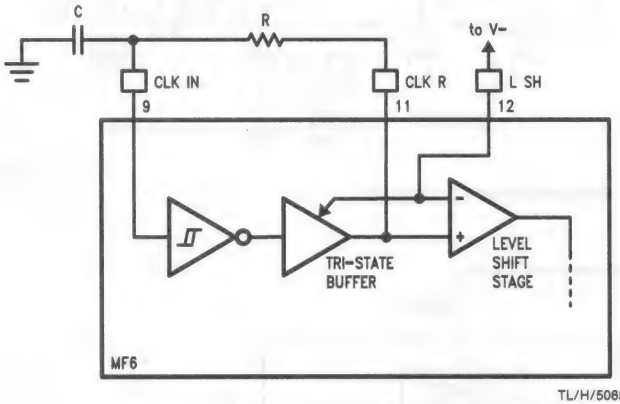
Pin	Description
L. Sh (12)	When the voltage level at this input exceeds [25%($V^+ - V^-$) + V^-] the internal tri-state buffer is disabled allowing the CLK R pin to become the clock input for the internal clock level shift stage. The CLK R threshold level is now 2V above the voltage applied to the L. Sh pin. Driving the CLK R pin with TTL logic levels can be accomplished through the use of split supplies and by tying the L. Sh pin to system ground.

1.0 MF6 Application Hints

The MF6 is comprised of a non-inverting unity gain lowpass sixth order Butterworth switched capacitor filter section and two undedicated CMOS Op-Amps. The switched capacitor topology makes the cutoff frequency (where the gain drops 3.01 dB below the DC gain) a direct ratio (100:1 or 50:1) of the clock frequency supplied to the lowpass filter. Internal integrator time constants set the filter's cutoff frequency. The resistive element of these integrators is actually a capacitor which is "switched" at the clock frequency (for a detailed discussion see Input Impedance Section). Varying the clock frequency changes the value of this resistive element and thus the time constant of the integrators. The clock to cutoff frequency ratio (f_{CLK}/f_c) is set by the ratio of the input and feedback capacitors in the integrators. The higher the clock to cutoff frequency ratio (or the sampling rate) the closer this approximation is to the theoretical Butterworth response. The MF6 is available in f_{CLK}/f_c ratios of 50:1 (MF6-50) or 100:1 (MF6-100).

1.1 CLOCK INPUTS

The MF6 has a Schmitt-trigger inverting buffer which can be used to construct a simple R/C oscillator. The oscillator's frequency is dependent on the buffer's threshold levels as well as on the resistor/capacitor tolerance (see *Figure 1*).



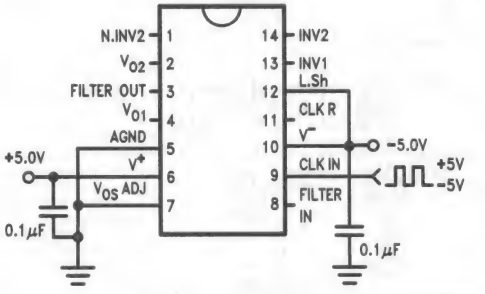
TL/H/5065–12

$$f_{CLK} = \frac{1}{RC \ln\left[\left(\frac{V_{CC} - V_{T-}}{V_{CC} - V_{T+}}\right)\frac{V_{T+}}{V_{T-}}\right]}$$

Typically for $V_{CC} = V^+ - V^- = 10V$:

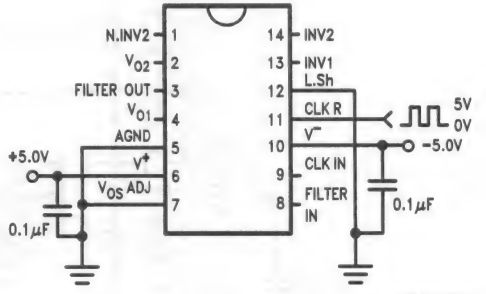
$$f_{CLK} = \frac{1}{1.69\ RC}$$

FIGURE 1. Schmitt Trigger R/C Oscillator



TL/H/5065–3

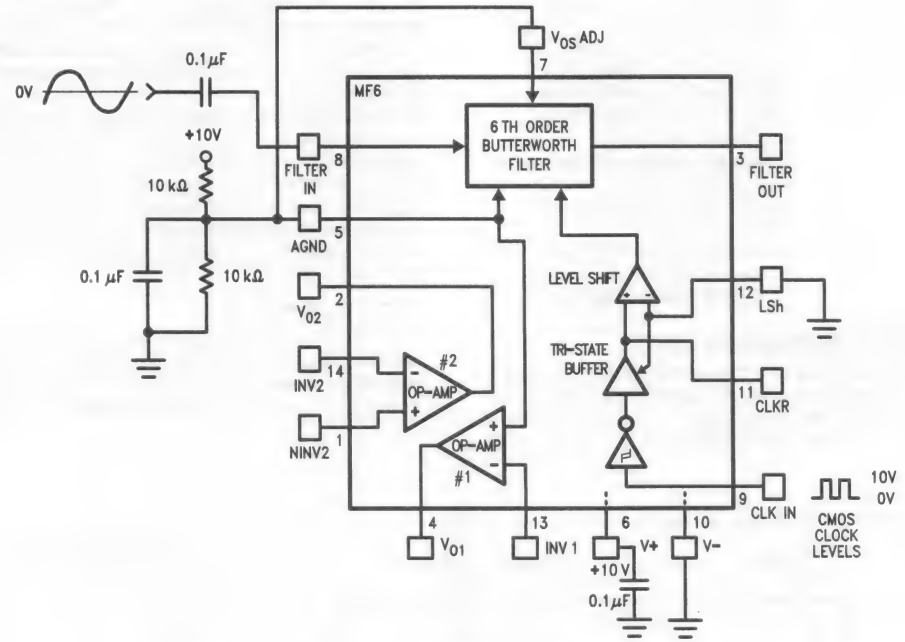
FIGURE 2. Dual Supply Operation MF6 Driven with CMOS Logic Level Clock ($V_{IH} \geq 0.8\ V_{CC}$ and $V_{IL} \leq 0.2\ V_{CC}$ where $V_{CC} = V^+ - V^-$)



TL/H/5065–4

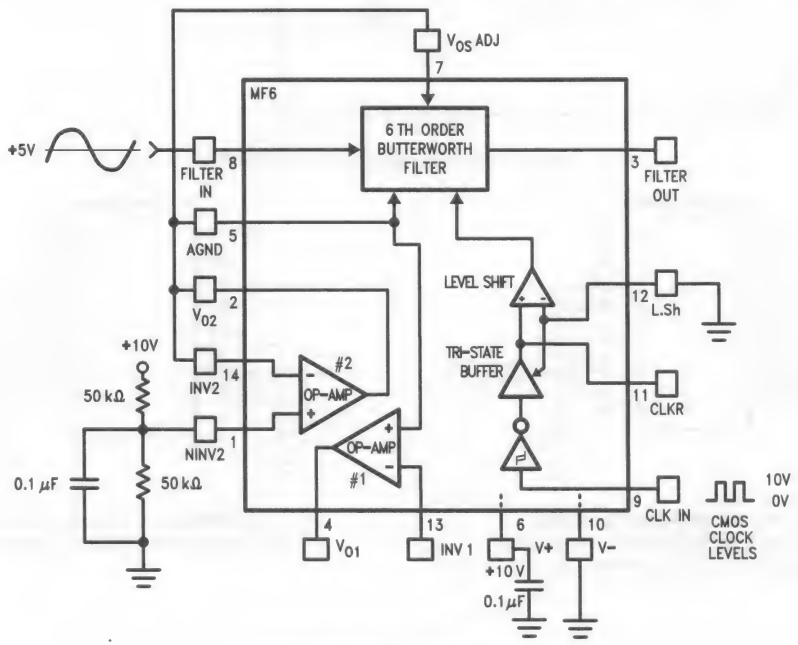
FIGURE 3. Dual Supply Operation MF6 Driven with TTL Logic Level Clock

Application Hints (Continued)



TL/H/5065–14

a) Resistor Biasing of AGND

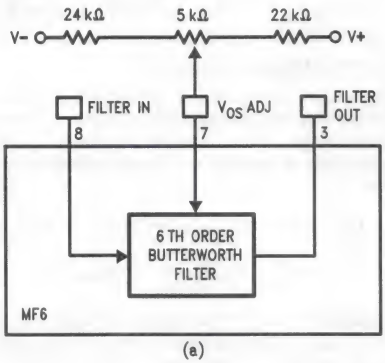


TL/H/5065–15

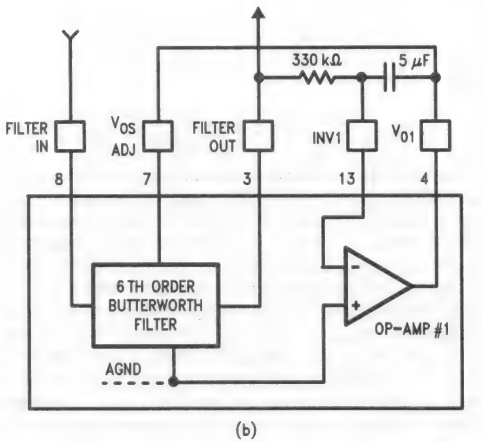
b) Using Op-Amp 2 to Buffer AGND

FIGURE 4. Single Supply Operation

Application Hints (Continued)



TL/H/5065–16



TL/H/5065–17

FIGURE 5. V_{OS} Adjust Schemes

Schmitt-trigger threshold voltage levels can change significantly causing the R/C oscillator's frequency to vary greatly from part to part.

Where accuracy in f_C is required an external clock can be used to drive the CLK R input of the MF6. This input is TTL logic level compatible and also presents a very light load to the external clock source (~2 μA) with split supplies and L. Sh tied to system ground. The logic level is programmed by the voltage applied to level shift (L. Sh) pin (See the Pin description for L. Sh pin).

1.2 POWER SUPPLY BIASING

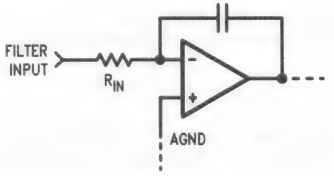
The MF6 can be biased from a single supply or dual split supplies. The split supply mode shown in *Figures 2* and *3* is the most flexible and easiest to implement. As discussed earlier split supplies, ±5V to ±7V, will enable the use of TTL or CMOS clock logic levels. *Figure 4* shows two schemes for single supply biasing. In this mode only CMOS clock logic levels can be used.

1.3 OFFSET ADJUST

The VosADJ pin is used in adjusting the output offset level of the filter section. If this pin is not used it must be tied to the analog ground (AGND) level, either mid-supply for single ended supply operation or ground for split supply operation. This pin sets the zero reference for the output of the filter. The implementation of this pin can be seen in *Figure 5*. In 5(a), DC offset is adjusted using a potentiometer; in 5(b), the Op-Amp integrator circuit keeps the average DC output level at AGND. The circuit in 5(b) is therefore appropriate only for AC-coupled signals and signals biased at AGND.

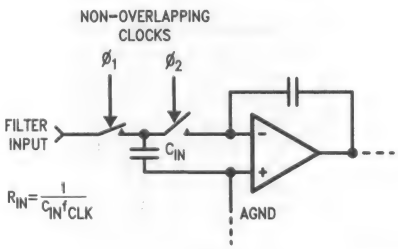
1.4 INPUT IMPEDANCE

The MF6 lowpass filter input (FILTER IN pin) is not a high impedance buffer input. This input is a switched capacitor resistor equivalent, and its effective impedance is inversely proportional to the clock frequency. The equivalent circuit of the input to the filter can be seen in *Figure 6*. The input capacitor charges to the input voltage (V_{in}) during one half of the clock period, during the second half the charge is



TL/H/5065–18

a) Equivalent Circuit for MF6 Filter Input



TL/H/5065–19

b) Actual Circuit for MF6 Filter Input

FIGURE 6. MF6 Filter Input

transferred to the feedback capacitor. The total transfer of charge in one clock cycle is therefore $Q = C_{in}V_{in}$, and since current is defined as the flow of charge per unit time the average input current becomes

$$I_{in} = Q/T$$

(where T equals one clock period) or

$$I_{in} = \frac{C_{in}V_{in}}{T} = C_{in}V_{in}f_{CLK}$$

The equivalent input resistor (R_{in}) then can be defined as

$$R_{in} = V_{in}/I_{in} = \frac{1}{C_{in}f_{CLK}}$$

The input capacitor is 2 pF for the MF6-50 and 1 pF for the

Application Hints (Continued)

MF6-100, so for the MF6-100

$$R_{in} = \frac{1 \times 10^{12}}{f_{CLK}} = \frac{1 \times 10^{12}}{f_c \times 100} = \frac{1 \times 10^{10}}{f_c}$$

and

$$R_{in} = \frac{5 \times 10^{11}}{f_{CLK}} = \frac{5 \times 10^{11}}{f_c \times 50} = \frac{1 \times 10^{10}}{f_c}$$

for the MF6-50. As shown in the above equations for a given cutoff frequency (f_c) the input impedance remains the same for the MF6-50 and the MF6-100. The higher the clock to center frequency ratio, the greater equivalent input resistance for a given clock frequency. As the cutoff frequency increases the equivalent input impedance decreases. This input resistance will form a voltage divider with the source impedance (R_{source}). Since R_{in} is inversely proportional to the cutoff frequency, operation at higher cutoff frequencies will be more likely to load the input signal which would appear as an overall decrease in gain to the output of the filter. Since the filter's ideal gain is unity its overall gain is given by:

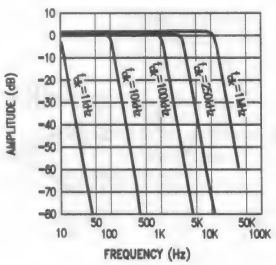
$$A_v = \frac{R_{in}}{R_{in} + R_{source}}$$

If the MF6-50 or the MF6-100 were set up for a cutoff frequency of 10 kHz the input impedance would be:

$$R_{in} = \frac{1 \times 10^{10}}{10\ kHz} = 1\ M\Omega$$

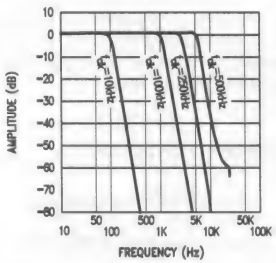
In this example with a source impedance of 10k the overall gain, if the MF6 had an ideal gain of 1 or 0 dB, would be:

$$A_v = \frac{1\ M\Omega}{10\ k\Omega + 1\ M\Omega} = 0.99009 \text{ or } -86.4\ mdB$$



TL/H/5065–20

FIGURE 7a. MF6-100 ±5V Supplies Amplitude Response



TL/H/5065–22

FIGURE 7c. MF6-100 ±2.5V Supplies Amplitude Response

Since the maximum overall gain error for the MF6 is ±0.3 dB with a $R_s \leq 2\ k\Omega$ the actual gain error for this case would be +0.21 dB to −0.39 dB.

1.5 CUTOFF FREQUENCY RANGE

The filter's cutoff frequency (f_c) has a lower limit caused by leakage currents through the internal switches discharging the stored charge on the capacitors. At lower clock frequencies these leakage currents can cause millivolts of error, for example:

$$f_{CLK} = 100\ Hz,\ I_{leakage} = 1\ pA,\ C = 1\ pF$$

$$V = \frac{1\ pA}{1\ pF\ (100\ Hz)} = 10\ mV$$

The propagation delay in the logic and the settling time required to acquire a new voltage level on the capacitors increases as the MF6 power supply voltage decreases. This causes a shift in the f_{CLK}/f_c ratio which will become noticeable when the clock frequency exceeds 250 kHz. The amplitude characteristic will stay within tolerance until f_{CLK} exceeds 500 kHz and will peak at about 0.5 dB at the corner frequency with a 1 MHz clock. The response of the MF6 is still a reasonable approximation of the ideal Butterworth lowpass characteristic as can be seen in *Figure 7*.

2.0 Designing with the MF6

Given any lowpass filter specification two equations will come in handy in trying to determine whether the MF6 will do the job. The first equation determines the order of the lowpass filter required:

$$n = \frac{\log\left(10^{0.1\,A_{min}} - 1\right) - \log\left(10^{0.1\,A_{max}} - 1\right)}{2 \log\left(f_s/f_b\right)} \qquad (1)$$

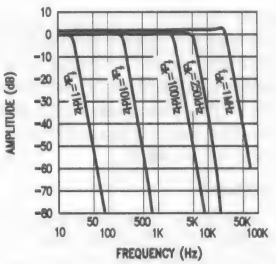


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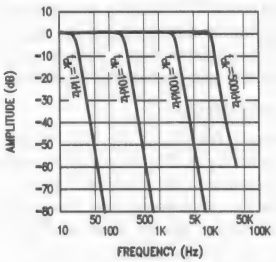
FIGURE 7b. MF6-50 ±5V Supplies Amplitude Response



TL/H/5065–23

FIGURE 7d. MF6-50 ±2.5V Supplies Amplitude Response

Designing with the MF6 (Continued)

where n is the order of the filter, A_{min} is the minimum stopband attenuation (in dB) desired at frequency f_s, and A_{max} is the passband ripple or attenuation (in dB) at frequency f_b. If the result of this equation is greater than 6, then more than a single MF6 is required.

The attenuation at any frequency can be found by the following equation:

$$\text{Attn}(f) = 10 \log [1 + (10^{0.1 A_{max}} - 1)\,(f/f_b)^{2n}] \text{ dB} \qquad (2)$$

where n = 6 (the order of the filter).

2.1 A LOWPASS DESIGN EXAMPLE

Suppose the amplitude response specification in *Figure 8* is given. Can the MF6 be used? The order of the Butterworth approximation will have to be determined using eq. 1:

A_{min} = 30 dB, A_{max} = 1.0 dB, f_s = 2 kHz, and f_b = 1 kHz

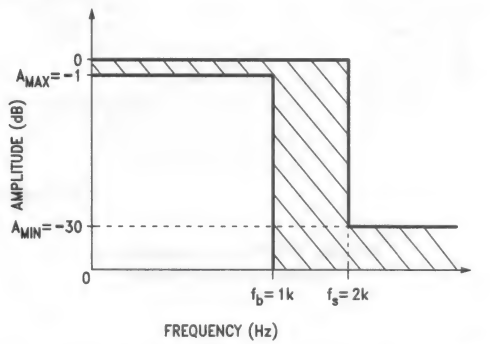
$$n = \frac{\log (10^3 - 1) - \log(10^{0.1} - 1)}{2 \log(2)} = 5.96$$

Since n can only take on integer values, n = 6. Therefore the MF6 can be used. In general, if n is 6 or less a single MF6 stage can be utilized.

Likewise, the attenuation at f_s can be found using equation 2 with the above values and n = 6 giving:

$$\text{Atten (2 kHz)} = 10 \log [1 + (10^{0.1} - 1)\,(2 \text{ kHz}/1 \text{ kHz})^{12}]$$
$$= 30.26 \text{ dB}$$

This result also meets the design specification given in *Figure 8* again verifying that a single MF6 section will be adequate.



TL/H/5065–24

FIGURE 8. Design Example Magnitude Response Specification Where the Response of the Filter Design Must Fall Within the Shaded Area of the Specification

Since the MF6's cutoff frequency f_c, which corresponds to a gain attenuation of −3.01 dB, was not specified in this example it needs to be calculated. Solving equation 2 where f = f_c as follows:

$$f_c = f_b \left[\frac{(10^{0.1(3.01 \text{ dB})} - 1)}{(10^{0.1\,A_{max}} - 1)}\right]^{1/(2n)}$$
$$= 1 \text{ kHz} \left[\frac{10^{0.301} - 1}{10^{0.1} - 1}\right]^{1/12}$$
$$= 1.119 \text{ kHz}$$

where $f_c = f_{CLK}/50$ or $f_{CLK}/100$.

To implement this example for the MF6-50 the clock frequency will have to be set to f_{CLK} = 50(1.116 kHz) = 55.8 kHz or for the MF6-100 f_{CLK} = 100(1.116 kHz) = 111.6 kHz.

2.2 CASCADING MF6s

In the case where a steeper stopband attenuation rate is required two MF6's can be cascaded (*Figure 9*) yielding a 12th order slope of 72 dB per octave. Because the MF6 is a Butterworth filter and therefore has no ripple in its passband, when MF6s are cascaded the resulting filter also has no ripple in its passband. Likewise the DC and passband gains will remain at 1V/V. The resulting response is shown in *Figure 10*.

In determining whether the cascaded MF6s will yield a filter that will meet a particular amplitude response specification, as above, equations 3 and 4 can be used, shown below.

$$n = \frac{\log (10^{0.05\,A_{min}} - 1) - \log (10^{0.05\,A_{max}} - 1)}{2 \log (f_s/f_b)} \qquad (3)$$

$$\text{Attn}(f) = 10 \log [1 + (10^{0.05\,A_{max}} - 1)\,(f/f_b)^{2n}] \text{ dB} \qquad (4)$$

where n = 6 (the order of each filter).

Equation 3 will determine whether the order of the filter is adequate (n ≤ 6) while equation 4 can determine if the required stopband attenuation is met and what actual cutoff frequency (f_c) is required to obtain the particular frequency response desired. The design procedure would be identical to the one shown in section 2.1.

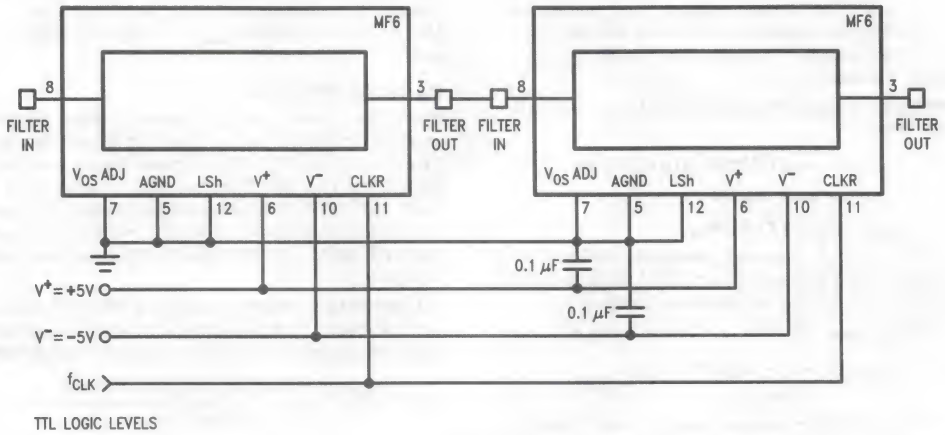
2.3 IMPLEMENTING A "NOTCH" FILTER WITH THE MF6

A "notch" filter with 60 dB of attenuation can be obtained by using one of the Op-Amps, available in the MF6, and three external resistors. The circuit and amplitude response are shown in *Figure 11*.

The frequency where the "notch" will occur is equal to the frequency at which the output signal of the MF6 will have the same magnitude but be 180 degrees out of phase with its input signal. For a sixth order Butterworth filter 180° phase shift occurs where $f = f_n = 0.742\,f_c$. The attenuation at this frequency is 0.12 dB which must be compensated for by making $R_1 = 1.014 \times R_2$.

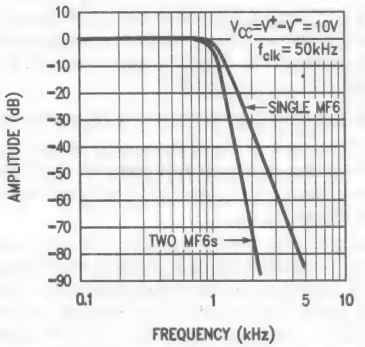
Since R_1 does not equal R_2 there will be a gain inequality above and below the notch frequency. At frequencies below the notch frequency ($f << f_n$), the signal through the filter has a gain of one and is non-inverting. Summing this with the input signal through the Op-Amp yields an overall gain of two or +6 dB. For $f >> f_n$, the signal at the output of the filter is greatly attenuated thus only the input signal will appear at the output of the Op-Amp. With $R_3 = R_1 = 1.014\,R_2$ the overall gain is 0.986 or −0.12 dB at frequencies above the notch.

Designing with the MF6 (Continued)



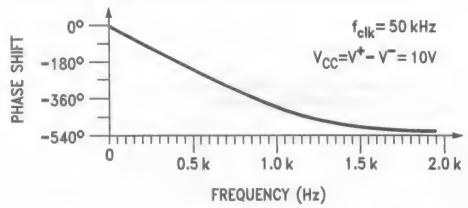
TL/H/5065–25

FIGURE 9. Cascading Two MF6s



TL/H/5065–26

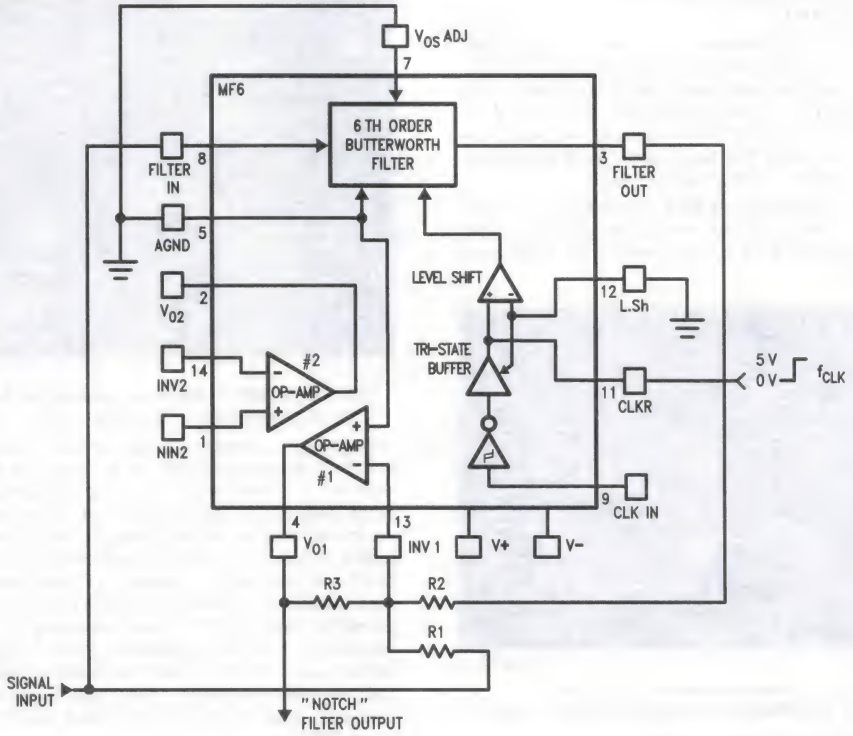
FIGURE 10a. One MF6-50 vs. Two MF6-50s Cascaded



TL/H/5065–27

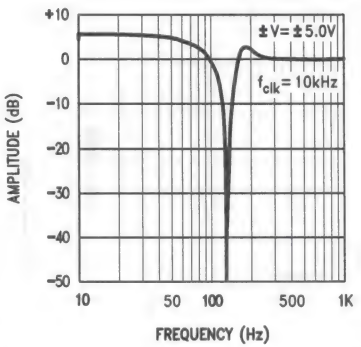
FIGURE 10b. Phase Response of Two Cascaded MF6-50s

Designing with the MF6 (Continued)



TL/H/5065–28

FIGURE 11a. "Notch" Filter



TL/H/5065–29

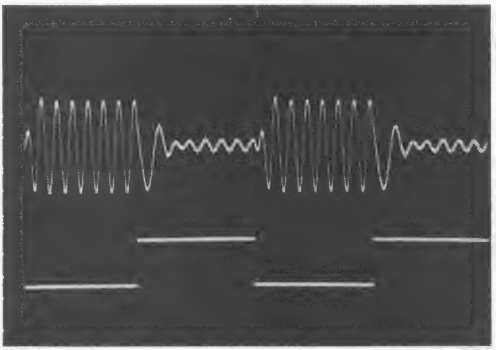
FIGURE 11b. MF6-50 "Notch" Filter Amplitude Response

Designing with the MF6 (Continued)

2.4 CHANGING CLOCK FREQUENCY INSTANTANEOUSLY

The MF6 will respond favorably to a sudden change in clock frequency. Distortion in the output signal occurs at the transition of the clock frequency and lasts approximately three cutoff frequency (f_c) cycles. As shown in *Figure 12,* if the control signal is low the MF6-50 has a 100 kHz clock making f_c = 2 kHz; when this signal goes high the clock frequency changes to 50 kHz yielding 1 kHz f_c.

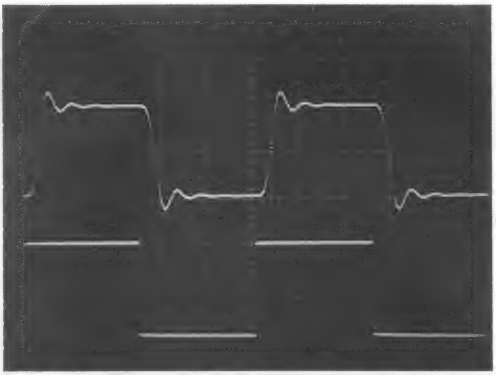
The transient response of the MF6 seen in *Figure 13* is also dependent on the f_c and thus the f_{CLK} applied to the filter. The MF6 responds as a classical sixth order Butterworth lowpass filter.



TL/H/5065–30

f_{IN} = 1.5 kHz (scope time base = 2 ms/div)

FIGURE 12. MF6-50 Abrupt Clock Frequency Change

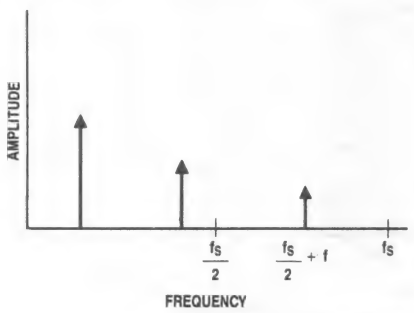


TL/H/5065–31

FIGURE 13. MF6-50 Step Input Response, Vertical = 2V/div., Horizontal = 1 ms/div., f_{CLK} = 100 kHz

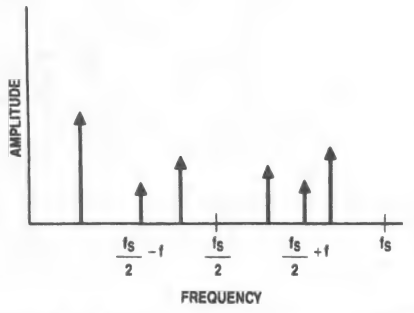
2.5 ALIASING CONSIDERATIONS

Aliasing effects have to be taken into consideration when input signal frequencies exceed half the sampling rate. For the MF6 this equals half the clock frequency (f_{CLK}). When the input signal contains a component at a frequency higher than half the clock frequency, as in *Figure 14a,* that component will be "reflected" about $f_{CLK}/2$ into the frequency range *below* $f_{CLK}/2$ as in *Figure 14b.* If this component is within the passband of the filter and of large enough amplitude it can cause problems. Therefore if frequency components in the input signal exceed $f_{CLK}/2$ they must be attenuated before being applied to the MF6 input. The necessary amount of attenuation will vary depending on system requirements. In critical applications the signal components above $f_{CLK}/2$ will have to be attenuated at least to the filter's residual noise level. An example circuit is shown in *Figure 15* using one of the uncommitted Op-Amps available in the MF6.



TL/H/5065–37

(a) Input Signal Spectrum

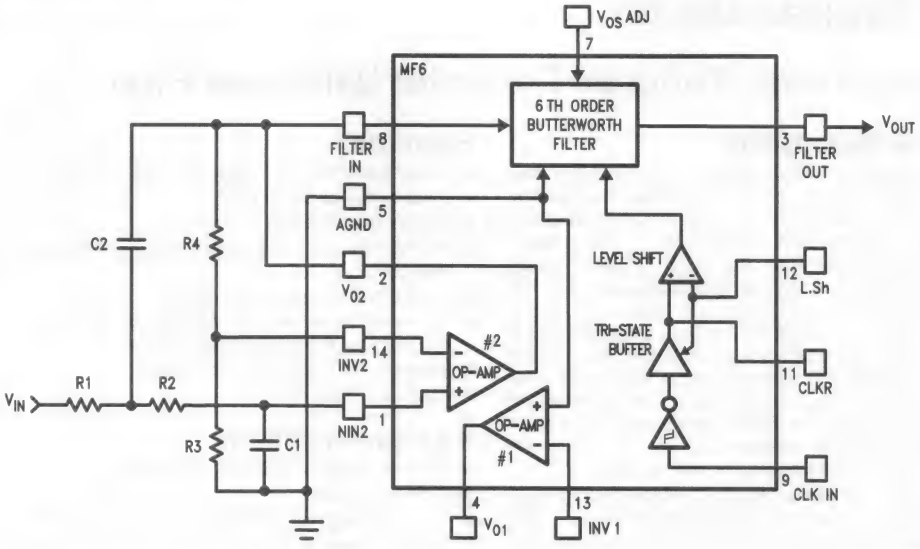


TL/H/5065–38

(b) Output Signal Spectrum. Note that the input signal at $f_S/2 + f$ causes an output signal to appear at $f_S/2 - f$.

Figure 14. The phenomenon of aliasing in sampled-data systems. An input signal whose frequency is greater than one-half the sampling frequency will cause an output to appear at a frequency lower than one-half the sampling frequency. In the MF6, $f_S = f_{CLK}$.

Designing with the MF6 (Continued)



TL/H/5065–34

$$f_0 = \frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}$$

$H_0 = R_4/R_3$ ($H_0 = 1$ when R_3 and R_4 are omitted and V_{O2} is directly tied to INV2).

Design Procedure:

pick C_1

$$R_2 = \frac{1}{2QC_1\omega_0}$$

for a 2nd Order Butterworth $Q = 0.707$

$$R_2 = \frac{0.113}{C_1f_0}$$

make $R_1 = R_2$

and

$$C_2 = \frac{1}{(2\pi f_0R_1)^2C_1}$$

Note: The parallel combination of R_4 (if used), R_1 and R_2 should be $\geq 10\ k\Omega$ in order not to load Op-Amp #2.

FIGURE 15. Second Order Butterworth Anti-Aliasing Filter Using Uncommitted Op-Amp #2

National Semiconductor

MF8 4th-Order Switched Capacitor Bandpass Filter

General Description

The MF8 consists of two second-order bandpass filter stages and an inverting operational amplifier. The two filter stages are identical and may be used as two tracking second-order bandpass filters, or cascaded to form a single fourth-order bandpass filter. The center frequency is controlled by an external clock for optimal accuracy, and may be set anywhere between 0.1 Hz and 20 kHz. The ratio of clock frequency to center frequency is programmable to 100:1 or 50:1. Two inputs are available for TTL or CMOS clock signals. The TTL input will accept logic levels referenced to either the negative power supply pin or the ground pin, allowing operation on single or split power supplies. The CMOS input is a Schmitt inverter which can be made to self-oscillate using an external resistor and capacitor.

By using the uncommitted amplifier and resistors for negative feedback, any all-pole (Butterworth, Chebyshev, etc.) filter can be formed. This requires only three resistors for a fourth-order bandpass filter. Q of the second-order stages may be programmed to any of 31 different values by the five "Q logic" pins. The available Q values span a range from 0.5 through 90. Overall filter bandwidth is programmed by connecting the appropriate Q logic pins to either V^+ or V^-. Filters with order higher than four can be built by cascading MF8s.

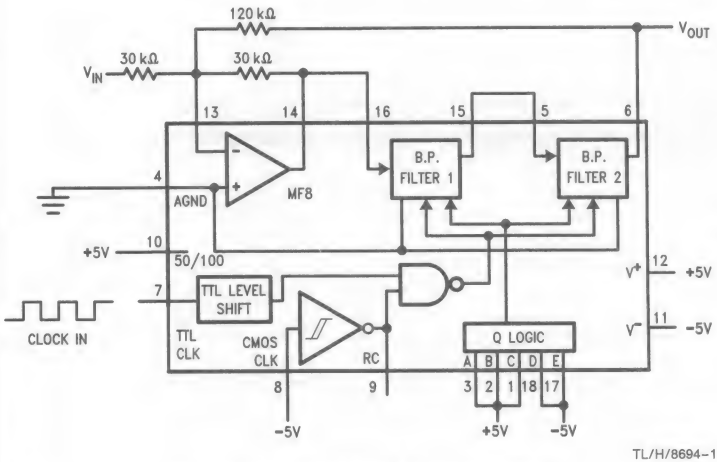
Features

- Center frequency set by external clock
- Q set by five-bit digital word
- Uncommitted inverting op amp
- 4th-order all-pole filters using only three external resistors
- Cascadable for higher-order filters
- Bandwidth, response characteristic, and center frequency independently programmable
- Separate TTL and CMOS clock inputs
- 18 pin 0.3" wide package

Key Specifications

- Center frequency range 0.1 Hz to 20 kHz
- Q range 0.5 to 90
- Supply voltage range 9V to 14V (±4.5V to ±7V)
- Center frequency accuracy 1% over full temperature range

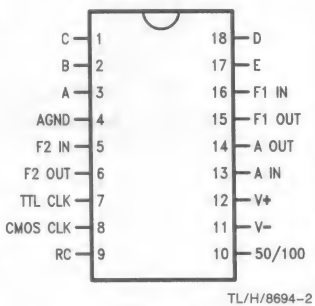
Typical Application & Connection Diagrams



TL/H/8694–1

Fourth-Order Butterworth Bandpass Filter

Dual-In-Line Package



TL/H/8694–2

Top View

Order Number MF8CCJ or MF8CCN
See NS Package Number J18A or N18A

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage ($V_S = V^+ - V^-$)	−0.3V to +15V
Voltage at any Input (Note 2)	V^- −0.3V to V^+ +0.3V
Input Current at any Input Pin (Note 2)	±1 mA
Output Short-Circuit Current (Note 7)	±1 mA
Power Dissipation (Note 3)	500 mW
Storage Temperature	−65°C to +150°C
Soldering Information:	
J Package: 10 sec.	260°C
N Package: 10 sec.	300°C
SO Package: Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

ESD rating is to be determined.

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

Operating Ratings (Note 1)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
MF8CCN	$0°C \le T_A \le +70°C$
MF8CCJ	$-40°C \le T_A \le +85°C$
Supply Voltage ($V_S = V^+ - V^-$)	+9V to +14V
$f_{CLK} \times Q$ Range	
for 10 Hz ≤ f_{CLK} ≤ 250 kHz	any Q
for 250 kHz ≤ f_{CLK} ≤ 1 MHz	$f_{CLK} \times Q \le 5$ MHz

Filter Electrical Characteristics

The following specifications apply for $V^+ = +5V$, $V^- = -5V$, C_{LOAD} = 50 pF and R_{LOAD} = 50 kΩ on filter output unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Symbol	Parameter (Notes 4, 5)	Conditions	MF8CCN Typical (Note 9)	MF8CCN Tested Limit (Note 10)	MF8CCN Design Limit (Note 11)	MF8CCJ Typical (Note 9)	MF8CCJ Tested Limit (Note 10)	MF8CCJ Design Limit (Note 11)	Units
H_o	Gain at f_o	f_{CLK} = 250 kHz, 100:1, ABCDE = 11100	6.02 ±.05	6.02 ±0.2		6.02 ±0.05	6.02 ±0.2		dB
Q	Q		3.92 ±2%	3.92 ±6%		3.92 ±2%	3.92 ±6%		
R	f_{CLK}/f_o		99.2 ±0.3%	99.2 ±1%		99.2 ±0.3%	99.2 ±1%		
H_o	Gain at f_o	f_{CLK} = 250 kHz, 100:1, ABCDE = 10011	6.02 ±0.2	6.02 ±0.5		6.02 ±0.2	6.02 ±0.5		dB
Q	Q		15.5 ±3%	15.5 ±8%		15.5 ±3%	15.5 ±8%		
R	f_{CLK}/f_o		99.7 ±0.3%	99.7 ±1%		99.7 ±0.3%	99.7±1%		
H_o	Gain at f_o	f_{CLK} = 250 kHz, 50:1, ABCDE = 00001	5.85 ±0.4	5.85 ±1		5.85 ±0.4	5.85 ±1		dB
Q	Q		55 ±5%	55 ±10%		55 ±5%	55 ±10%		
R	f_{CLK}/f_o		49.9 ±0.2%	49.9 ±1%		49.9 ±0.2%	49.9 ±1%		
H_o	Gain at f_o	V_S = ±5V ±5%, $f_{CLK} \le$ 250 kHz	6.02 ±0.5		**6.02 ±1.5**	6.02 ±0.5		**6.02 ±1.5**	dB
$\Delta Q/Q_{TH}$	Q Deviation from Theoretical (See Table I)	V_S = ±5V ±5%, $f_{CLK} \le$ 250 kHz, Q > 1	±5%		**±15%**	±5%		**±15%**	
		$f_{CLK} \le$ 100 kHz, 1 < Q < 57	±2%		**±6%**	±2%		**±6%**	
$\Delta R/R_{TH}$	f_{CLK}/f_o Deviation from Theoretical (See Table I)	V_S = ±5V ±5%, $f_{CLK} \le$ 250 kHz	±0.3%		**±1%**	±0.3%		**±1%**	
Q	Q	f_{CLK} = 250 kHz, 50:1, ABCDE = 00110	10.6 ±2%		**10.6 ±6%**	10.6 ±2%	**10.6 ±8%**		
	Dynamic Range (Note 6)	ABCDE = 11100	86			86			dB
		ABCDE = 10011	80			80			dB
		ABCDE = 00001	75			75			dB
	Clock Feedthrough	Filter and Op Amp, $f_{CLK} \le$ 250 kHz, Q ≤ 1	80			80			mV
		Q > 1	40			40			mV
I_S	Maximum Supply Current	f_{CLK} = 250 kHz, no loads on outputs	9	12	**12**	9	**13**		mA
V_{OS}	Maximum Filter Output Offset Voltage	f_{CLK} = 250 kHz, Q = 4, 50:1	±40	±120		±40	±120		mV
		100:1	±80	±240		±80	±240		mV
V_{OUT}	Minimum Filter Output Swing	R_{LOAD} = 5 kΩ (Note 6)	±4.1	±3.8	**±3.8**	±4.1	**±3.6**		V

Op Amp Electrical Characteristics

The following specifications apply for $V^+ = +5V$, $V^- = -5V$ and no load on the Op Amp output unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Symbol	Parameter	Conditions	MF8CCN			MF8CCJ			Units
			Typical (Note 9)	Tested Limit (Note 10)	Design Limit (Note 11)	Typical (Note 9)	Tested Limit (Note 10)	Design Limit (Note 11)	
V_{OS}	Maximum Input Offset Voltage		±8	±20		±8	±20		mV
I_B	Maximum Input Bias Current		10			10			pA
V_{OUT}	Minimum Output Voltage Swing	$R_{LOAD} = 5\ k\Omega$	±3.8	±3.5	**±3.4**	±3.8	**±3.1**		V
A_{VOL}	Open Loop Gain		80			80			dB
GBW	Gain Bandwidth Product		1.8			1.8			MHz
SR	Slew Rate		10			10			V/μs

Logic Input and Output Characteristics

The following specifications apply for $V^+ = +10V$ and $V^- = 0V$ unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Symbol	Parameter		Conditions	MF8CCN			MF8CCJ			Units
				Typical (Note 9)	Tested Limit (Note 10	Design Limit (Note 11)	Typical (Note 9)	Tested Limit (Note 10)	Design Limit (Note 11)	
V_T^+	Positive Threshold Voltage on pin 8	Min	$V_S = V^+ - V^-$ referred to $V^- = 0V$ (Note 8)	$0.7V_S$	$0.58V_S$		$0.7V_S$	$0.58V_S$		V
		Max		$0.7V_S$	$0.89V_S$		$0.7V_S$	$0.89V_S$		V
V_T^-	Negative Threshold Voltage on pin 8	Min	$V_S = V^+ - V^-$ referred to $V^- = 0V$ (Note 8)	$0.35V_S$	$0.11V_S$		$0.35V_S$	$0.11V_S$		V
		Max		$0.35V_S$	$0.47V_S$		$0.35V_S$	$0.47V_S$		V
V_{OH}	Output Voltage on pin 9 (Note 12)	Min High	$I_O = -10\ \mu A$		9.0	**9.0**		**9.0**		V
V_{OL}		Max Low	$I_O = +10\ \mu A$		1.0	**1.0**		**1.0**		V
I_{OH}	Output Current on pin 9	Min Source	Pin 9 tied to V^-	6.0	3.0		6.0	3.0		mA
I_{OL}		Min Sink	Pin 9 tied to V^+	5.0	2.5		5.0	2.5		mA
V_{IH}	Input Voltage on pins: 1, 2, 3, 10, 17, & 18 (Note 12)	Min High		7.0		**9.0**	7.0	**9.0**		V
V_{IL}		Max Low		3.0		**1.0**	3.0	**1.0**		V
I_{IN}	Input Current on pins: 1, 2, 3, 7, 8, 10, 17, & 18				10	**10**		**10**		μA
V_{IH}	Input Voltage on pin 7	Min High	$V^+ = +10V$, $V^- = 0V$ or $V^+ = +5V$, $V^- = -5V$		2.0	**2.0**		**2.0**		V
V_{IL}		Max Low			0.8	**0.8**		**0.8**		V

Note 1: Absolute Maximum Raings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: When the applied voltage at any pin falls outside the power supply voltages ($V_{IN} < V^-$ or $V_{IN} > V^+$), the absolute value of current at that pin should be limited to 1 mA or less.

Note 3: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, Θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\Theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 125°C$, and the typical junction-to-ambient thermal resistance of the MF8CCN when board mounted is 50°C/W. For the MF8CCJ, this number increases to 65°C/W.

Note 4: The center frequency of each 2nd-order filter section is defined as the frequency where the phase shift through the filter is zero.

Note 5: Q is defined as the measured center frequency divided by the measured bandwidth, where the bandwidth is the difference between the two frequencies where the gain is 3 dB less than the gain measured at the center frequency.

Note 6: Dynamic range is defined as the ratio of the tested minimum output swing of 2.69 Vrms (±3.8V peak-to-peak) to the wideband noise over a 20 kHz bandwidth. For Qs of 1 or less the dynamic range and output swing will degrade because the gain at an internal node is 2/Q. Keeping the input signal level below 1.23xQ Vrms will avoid distortion in this case.

Note 7: If it is possible for a signal output (pin 6, 14, or 15) to be shorted to V^+, V^- or ground, add a series resistor to limit output current.

Note 8: If V^- is anything other than 0V then the value of V^- should be added to the values given in the table. For example for $V^+ = +5V$ and $V^- = -5V$ the typical $V_T^+ = 0.7\ (10V) + (-5V) = +2V$.

Note 9: Typicals are at 25°C and represent the most likely parametric norm.

Note 10: Tested Limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 11: Design Limits are guaranteed but not 100% tested. These limits are not used to calculate outgoing quality levels.

Note 12: These logic levels have been referenced to V^-. The logic levels will shift accordingly for split supplies.

Pin Descriptions

Q Logic Inputs A, B, C, D, E (3, 2, 1, 18, 17): These inputs program the Qs of the two 2nd-order bandpass filter stages. Logic "1" is V^+ and logic "0" is V^-.

AGND (4): This is the analog and digital ground pin and should be connected to the system ground for split supply operation or biased to mid-supply for single supply operation. For best filter performance, the ground line should be "clean".

V^+ (12), V^- (11): These are the positive and negative power supply inputs. Decoupling the power supply pins with 0.1 μF or larger capacitors is highly recommended.

F1 IN (16), F2 IN (5): These are the inputs to the bandpass filter stages. To minimize gain error the source impedance should be less than 2 kΩ. Input signals should be referenced to AGND.

F1 OUT (15), F2 OUT (6): These are the outputs of the bandpass filter stages.

A IN (13): This is the inverting input to the uncommitted operational amplifier. The non-inverting input is internally connected to AGND.

A OUT (14): This is the output of the uncommitted operational amplifier.

50/100 (10): This pin sets the ratio of the clock frequency to the bandpass center frequency. Connecting this pin to V^+ sets the ratio to 100:1. Connecting it to V^- sets the ratio to 50:1.

TTL CLK (7): This is the TTL-level clock input pin. There are two logic threshold levels, so the MF8 can be operated on either single-ended or split supplies with the logic input referred to either V^- or AGND. When this pin is not used (or when CMOS logic levels are used), it should be connected to either V^+ or V^-.

CMOS CLK (8): This pin is the input to a CMOS Schmitt inverter. Clock signals with CMOS logic levels may be applied to this input. If the TTL input is used this pin should be connected to V^-.

RC (9): This pin allows the MF8 to generate its own clock signal. To do this, connect an external resistor between the RC pin and the CMOS Clock input, and an external capacitor from the CMOS Clock input to AGND. The TTL Clock input should be connected to V^- or V^+. When the MF8 is driven from an external clock, the RC pin should be left open.

1.0 Application Information

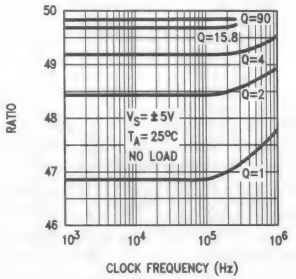
1.1 INTRODUCTION

A simplified block diagram for the MF8 is shown in *Figure 1*. The analog signal path components are two identical 2nd-order bandpass filters and an operational amplifier. Each filter has a fixed voltage gain of 2. The filters' cutoff frequency is proportional to the clock frequency, which may be applied to the chip from an external source or generated internally with the aid of an external resistor and capacitor. The proportionality constant f_{CLK}/f_0 can be set to either 50 or 100 depending on the logic level on pin 10. The "Q" of the two filters can have any of 31 values ranging from 0.5 to 90 and is set by the logic levels on pins 1, 2, 3, 17, and 18. Table I shows the available values of Q and the logic levels required to obtain them. The operational amplifier's non-inverting input is internally grounded, so it may be used only for inverting applications.

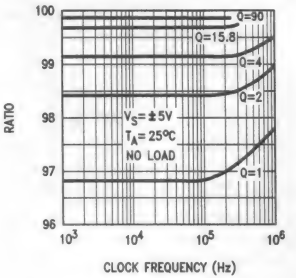
The components in the analog signal path can be interconnected in several ways, three of which are illustrated in *Figures 2a, 2b* and *2c*. The two second-order filter sections can be used as separate filters whose center frequencies track very closely as in *Figure 2a*. Each filter section has a high input impedance and low output impedance. The op amp may be used for gain scaling or other inverting functions. If sharper cutoff slopes are desired, the two filter sections may be cascaded as in *Figure 2b*. Again, the op amp is uncommitted. The circuit in *Figure 2c* uses both filter sections with the op amp and three resistors to build a "multiple feedback loop" filter. This configuration offers the greatest flexibility for fourth-order bandpass designs. Virtually any fourth-order all pole response shape (Butterworth, Chebyshev) can be obtained with a wide range of bandwidths, simply by proper choice of resistor values and Q. The three connection schemes in *Figure 2* will be discussed in more detail in Sections 1.4 and 1.5.

Typical Performance Characteristics

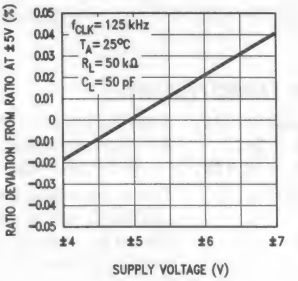
f_{CLK}/f_o Ratio vs Clock Frequency—50:1 Mode



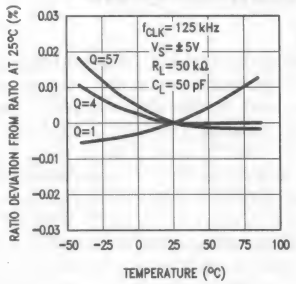
f_{CLK}/f_o Ratio vs Clock Frequency—100:1 Mode



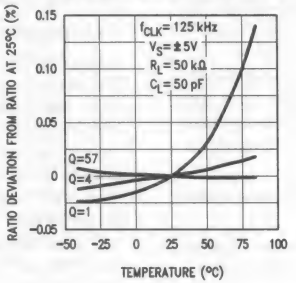
f_{CLK}/f_o Ratio vs Supply Voltage—50:1 and 100:1 Mode



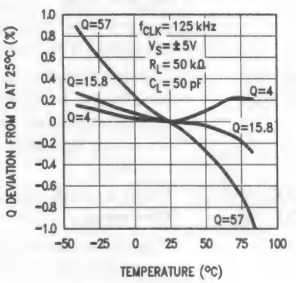
f_{CLK}/f_o Ratio vs Temperature—100:1 Mode



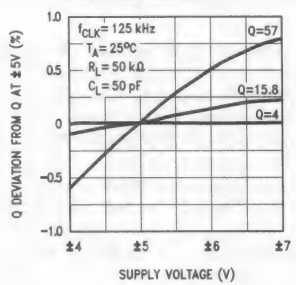
f_{CLK}/f_o Ratio vs Temperature—50:1 Mode



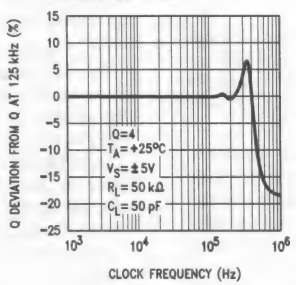
Q vs Temperature—50:1 and 100:1



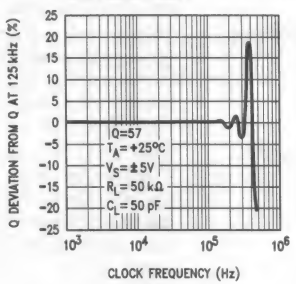
Q vs Supply Voltage—50:1 and 100:1



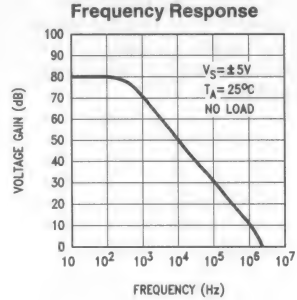
Q vs Clock Frequency—50:1 and 100:1



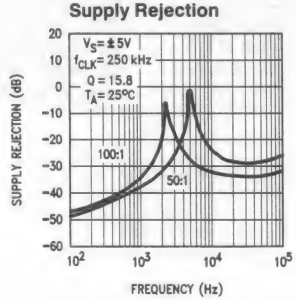
Q vs Clock Frequency—50:1 and 100:1



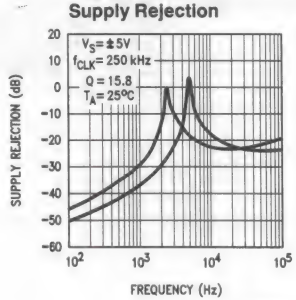
Op Amp—Open Loop Frequency Response



Positive Power Supply Rejection



Negative Power Supply Rejection



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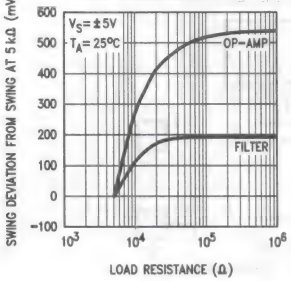
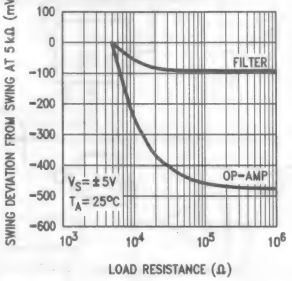
Typical Performance Characteristics (Continued)

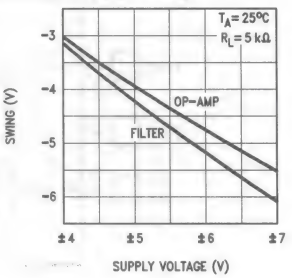
Positive Swing vs Load Resistance



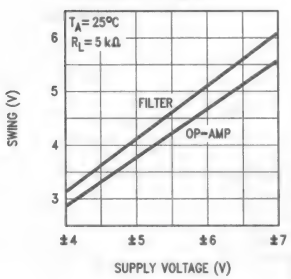
Negative Swing vs Load Resistance



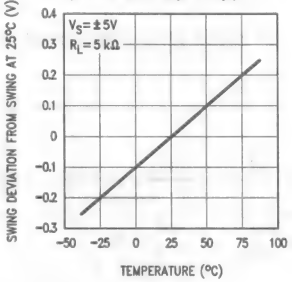
Negative Swing vs Supply Voltage



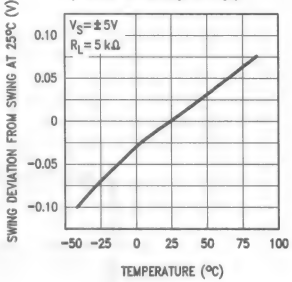
Positive Swing vs Supply Voltage



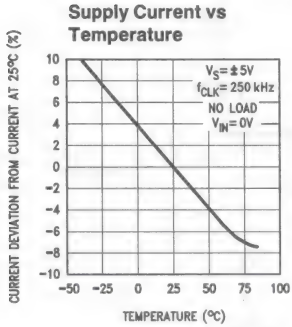
Negative Swing vs Temperature (Filter and Op Amp)



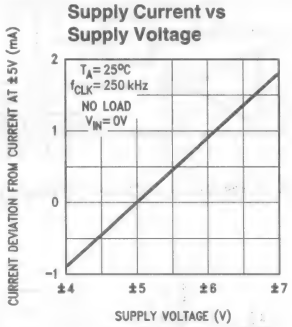
Positive Swing vs Temperature (Filter and Op Amp)



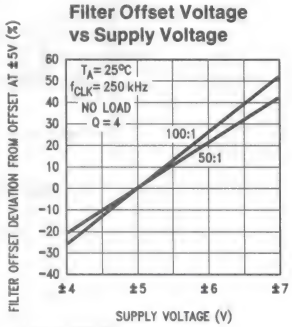
Supply Current vs Temperature



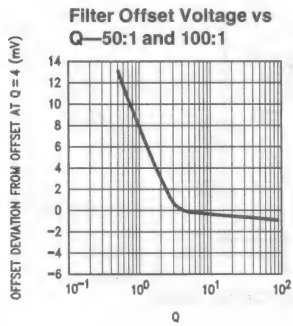
Supply Current vs Supply Voltage



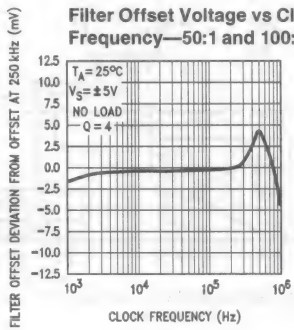
Filter Offset Voltage vs Supply Voltage



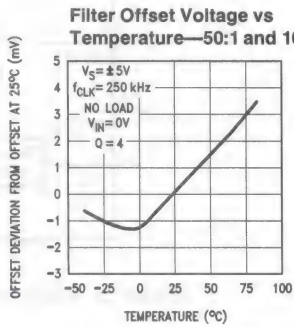
Filter Offset Voltage vs Q—50:1 and 100:1



Filter Offset Voltage vs Clock Frequency—50:1 and 100:1



Filter Offset Voltage vs Temperature—50:1 and 100:1



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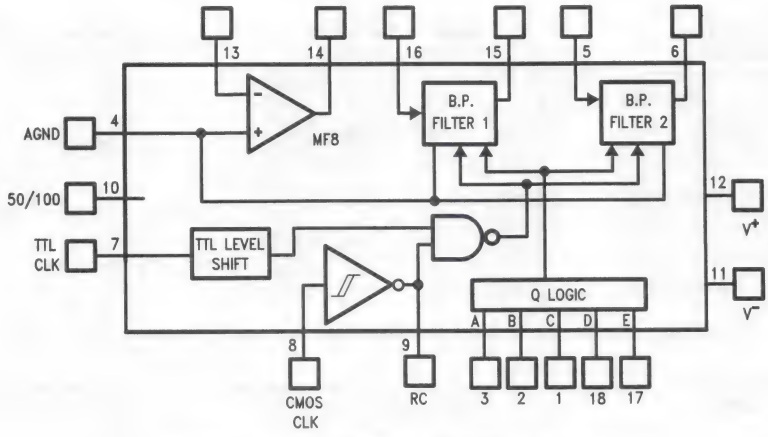
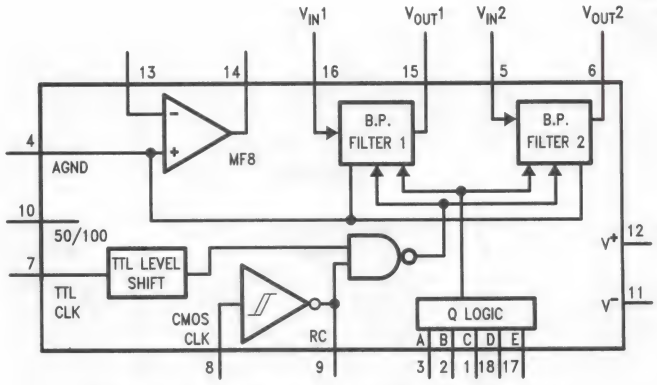
1.0 Application Information (Continued)



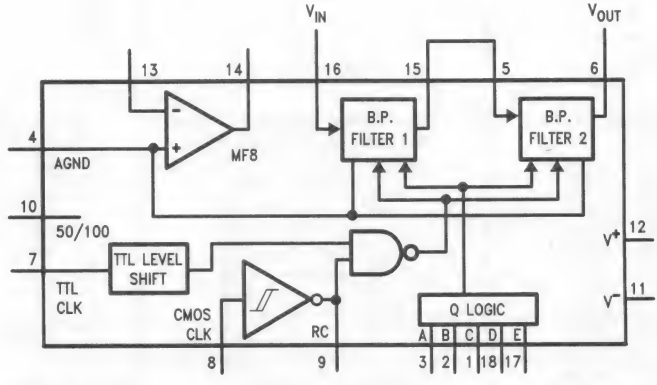
TL/H/8694–3

FIGURE 1. Simplified Block Diagram of the MF8



TL/H/8694–4

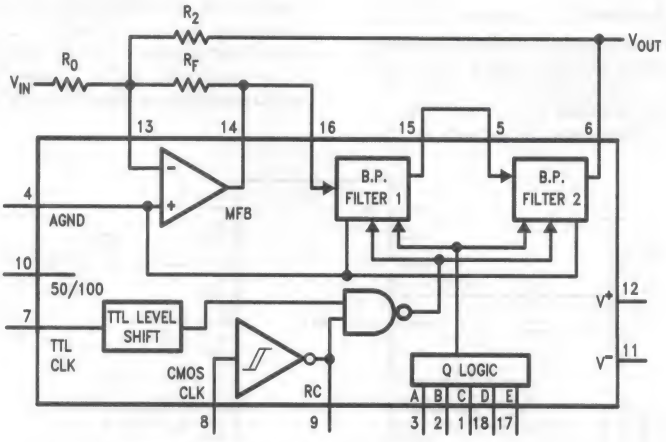
FIGURE 2a. Separate Second-Order "Tracking" Filters



TL/H/8694–5

FIGURE 2b. Fourth-Order Bandpass Made by Cascading Two Second-Order Stages

1.0 Application Information (Continued)



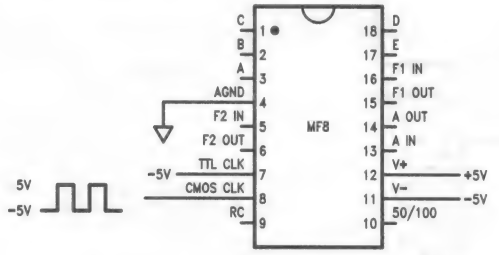
TL/H/8694–6

FIGURE 2c. Multiple Feedback Loop Connection

1.2 CLOCKS

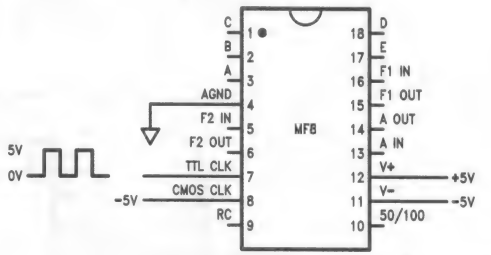
The MF8 has two clock input pins, one for CMOS logic levels and the other for TTL levels. The TTL (pin 7) input automatically adjusts its switching threshold to enable operation on either single or split power supplies. When this input is used, the CMOS logic input should be connected to pin 11(V^-). The CMOS Schmitt trigger input at pin 8 accepts CMOS logic levels. When it is used, the TTL input should be connected to either pin 11 (V^-) or pin 12 (V^+). The basic clock hookups for single and split supply operation are shown in *Figures 3* and *4*.

Clock signals derived from a crystal-controlled oscillator are recommended when maximum center frequency accuracy is desired, but in less critical applications the MF8 can generate its own clock signal as in *Figures 3c* and *4c*. An external resistor and capacitor determine the oscillation frequency. Tolerance of these components and part-to-part variations in Schmitt-trigger logic thresholds limit the accuracy of the RC clock frequency. In the self-clocked mode the TTL Clock input should be connected to either pin 11 or pin 12.



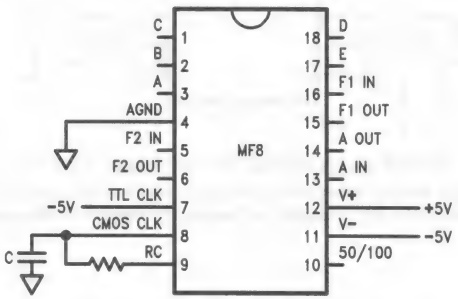
TL/H/8694–7

(a) MF8 Driven with CMOS Logic Level Clock



TL/H/8694–8

(b) MF8 Driven with TTL Logic Level Clock



$$f_{CLK} = \frac{1}{RC \ln\left|\left(\frac{V_S - V_{T-}}{V_S - V_{T+}}\right)\left(\frac{V_{T+}}{V_{T-}}\right)\right|}$$

Typically for $V_S^* = 10V$

$$f_{CLK} = \frac{1}{1.69\,RC}$$

$^*V_S = V^+ - V^-$

TL/H/8694–9

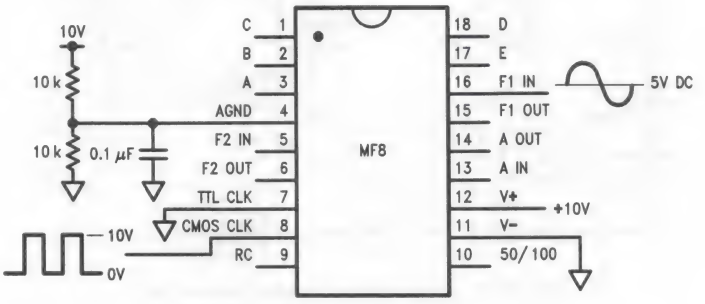
(c) MF8 Driven with Schmitt Trigger Oscillator

FIGURE 3. Dual Supply Operation

1.0 Application Information (Continued)

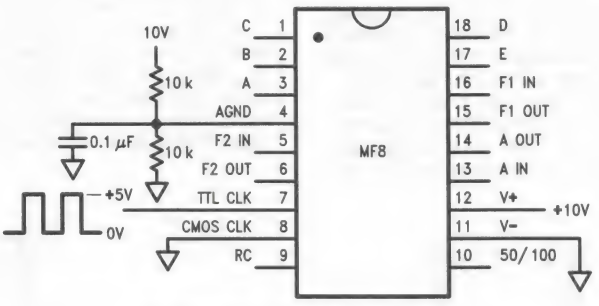
1.3 POWER SUPPLIES AND ANALOG GROUND

The MF8 can be operated from single or dual-polarity power supplies. For dual-supply operation, the analog ground (pin 4) should be connected to system ground. When single supplies are used, pin 4 should be biased to $V^+/2$ as in *Figures 3* and *4*. The input signal should either be capacitively coupled to the filter input or biased to $V^+/2$. It is strongly recommended that each power supply pin be bypassed to ground with at least a 0.1 μF ceramic capacitor. In single supply applications, with V^- connected to ground, V^+ and AGND should be bypassed to system ground.



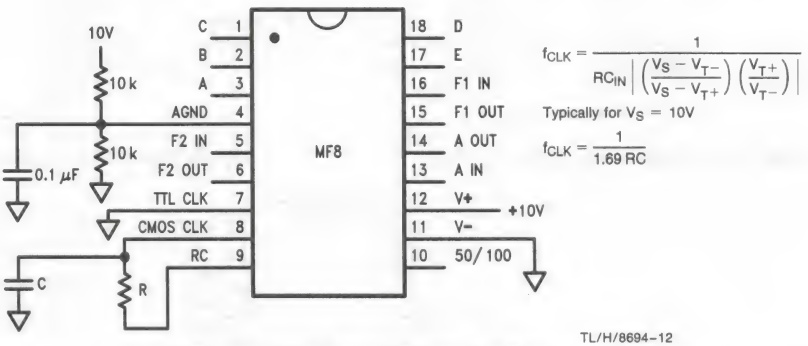
TL/H/8694–10

(a) MF8 Driven with CMOS Logic Level Clock



TL/H/8694–11

(b) MF8 Driven with TTL Logic Clock



TL/H/8694–12

(c) MF8 Driven with the Schmitt Trigger Oscillator

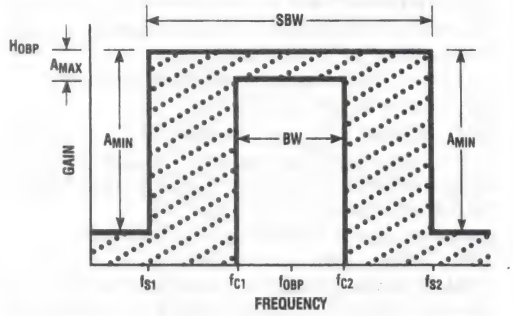
FIGURE 4. Single supply operation. The AGND pin must be biased to mid-supply. The input signal should be dc biased to mid-supply or capacitor-coupled to the input pin.

1.0 Application Information (Continued)

1.4 MULTIPLE FEEDBACK LOOP CONFIGURATION

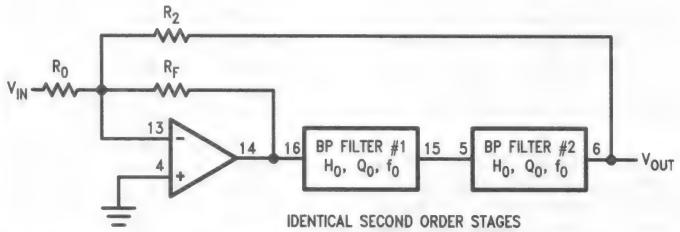
The multi-loop approach to building bandpass filters is highly flexible and stable, yet uses few external components. *Figure 5* shows the MF8's internal operational amplifier and two second-order filter stages with three external resistors in a fourth-order multiple feedback configuration. Higher-order filters may be built by adding more second-order sections and feedback resistors as in *Figure 6*. The filter's response is determined by the clock frequency, the clock-to-center-frequency ratio, the ratios of the feedback resistor values, and the Qs of the second-order filter sections. The design procedure for multiple feedback filters can be broken down into a few simple steps:

1) Determine the characteristics of the desired filter. This will depend on the requirements of the particular application. For a given application, the required bandpass response can be shown graphically as in *Figure 7*, which shows the limits for the filter response. *Figure 7* also makes use of several parameters that must be known in order to design a filter. These parameters are defined below in terms of *Figure 7*.



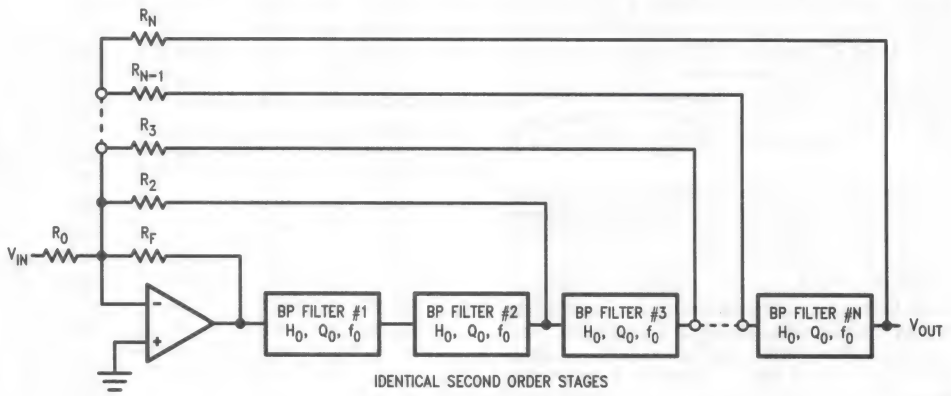
TL/H/8694–15

FIGURE 7. Graphical representation of the amplitude response specifications for a bandpass filter. The filter's response should fall within the shaded area.



TL/H/8694–13

FIGURE 5. General fourth-order multiple-feedback bandpass filter circuit. MF8 pin numbers are shown.



TL/H/8694–14

FIGURE 6. By adding more second-order filter stages and feedback resistors, higher order multiple-feedback filters may be built.

1.0 Application Information (Continued)

f_{C1} and f_{C2}: The filter's lower and upper cutoff frequencies. These define the filter's passband.

f_{S1} and f_{S2}: The boundaries of the filter's stopband.

BW: The filter's bandwidth. $BW = f_{C2} - f_{C1}$.

SBW: The width of the filter's stopband. $SBW = f_{S2} - f_{S1}$.

f_0: The center frequency of the filter. f_0 is equal to the geometric mean of f_{C1} and f_{C2}: $f_0 = \sqrt{f_{C1}f_{C2}}$. f_0 is also equal to the geometric mean of f_{S1} and f_{S2}.

H_{0BP}: The nominal passband gain of the bandpass filter. This is normally taken to be the gain at f_0.

f_0/BW: The ratio of the center frequency to the bandwidth. For second-order filters, this quantity is also known as "Q".

SBW/BW: The ratio of stopband width to bandwidth. This quantity is also called "Omega" and may be represented by the symbol "Ω".

A_{max}: The maximum allowable gain variation within the filter passband. This will depend on the system requirements, but typically ranges from a fraction of a dB to 3 dB.

A_{min}: The minimum allowable attenuation in the stopband. Again, the required value will depend on system constraints.

2). Choose a Butterworth or Chebyshev response characteristic. Butterworth bandpass filters are monotonic on either side of the center frequency, while Chebyshev filters will have "ripple" in the passband, but generally faster attenuation outside the passband. Chebyshev filters are specified according to the amount of ripple (in dB) within the passband.

3) Determine the filter order necessary to meet the response requirements defined above. This may be done with the aid of the nomographs in *Figures 8* and *9* for Butterworth and Chebyshev filters. To use the nomographs, draw a line through the desired values on the A_{MAX}/A_{MIN} scales to the left side of the graph. Draw a horizontal line to the right of this point and mark its intersection with the vertical line corresponding to the required ratio SBW/BW. The required filter order will be equal to the number of the curve falling on or just above the intersection of the two lines. This is illustrated in *Figure 10* for a Chebyshev filter with 1 dB ripple, 30 dB minimum attenuation in the stopband, and SBW/BW = 3. From the *Figure,* the required filter order is 6.

4) The design tables in section 2.0 can now be used to find the component values that will yield the desired response for filters of order 4 through 12. The "K_n" give the ratios of resistors "R_n" to R_F, and K_Q is Q divided by f_0/BW.

As an example of the Tables' use, consider a fourth-order Chebyshev filter with 0.5 dB ripple and $f_0/BW = 6$. Begin by choosing a convenient value for R_F, such as 100 kΩ. From the "0.5 dB Chebyshev" filter table, $K_0 = R_0/R_F = 1.3405$. This gives $R_0 = R_F \times 1.345 = 134.05k$. In a similar manner, R_2 is found to equal 201.61k. Q is found using the column labeled K_Q. This gives $Q = K_Q \times f_0/BW = 8.4174$.

Table I shows the available Q values; the nearest value is 8.5, which is programmed by tying pins 1, 2, 3, and 18 to V^+ and pin 17 to V^-.

Note that the resistor values obtained from the tables are normalized for center frequency gain $H_{OBP} = 1$. For different gains, simply divide R_0 by the desired gain.

5) Choose the clock-to-center-frequency ratio. This will nominally be 100:1 when pin 10 is connected to pin 12(V^+) and 50:1 when pin 10 is connected to pin 11(V^-). 100:1 generally gives a response curve nearer the ideal and fewer (if any) problems with aliasing, while 50:1 allows operation over the highest octave of center frequencies (10 kHz to 20 kHz). Supply the MF8 with a clock signal of the appropriate frequency to either the TTL or CMOS input, depending on the available clock logic levels.

TABLE I. Q and Clock-to-Center-Frequency Ratio Versus Logic Levels on "Q-set" Pins

	50:1 mode		100:1 mode	
ABCDE	**F_{CLK}/F_o**	**Q**	**F_{CLK}/F_o**	**Q**
10000	43.7	0.45	94.0	0.47
11000	45.8	0.71	95.8	0.73
01000	46.8	0.96	96.8	0.98
10100	48.4	2.0	98.4	2.0
00100	48.7	2.5	98.7	2.5
01100	48.9	3.0	98.9	3.0
11100	49.2	4.0	99.2	4.0
01010	49.3	5.0	99.3	5.0
10010	49.4	5.7	99.4	5.7
10110	49.4	6.4	99.4	6.4
00010	49.5	7.6	99.5	7.6
11110	49.6	8.5	99.6	8.5
00110	49.6	10.6	99.6	10.6
11001	49.6	11.7	99.6	11.7
11010	49.7	12.5	99.7	12.5
11101	49.7	13.6	99.7	13.6
01001	49.7	14.7	99.7	14.7
10011	49.7	15.8	99.7	15.8
10101	49.7	16.5	99.7	16.5
01110	49.7	17	99.7	17
10001	49.8	19	99.8	19
10111	49.8	22	99.8	22
11011	49.8	27	99.8	27
11111	49.8	30	99.8	30
00101	49.8	33	99.8	33
01011	49.8	40	99.8	40
00111	49.8	44	99.8	44
00001	49.9	57	99.9	57
01101	49.9	68	99.9	68
00011	49.9	79	99.9	79
01111	49.9	90	99.9	90

1.0 Application Information (Continued)

Higher-order filters are designed in a similar manner. An eighth-order Chebyshev with 0.1 dB ripple, center frequency equal to 1 kHz, and 100 Hz bandwidth, for example, could be built as in *Figure 11* with the following component values:

$R_0 = 79.86k$
$R_F = 100k$
$R_2 = 57.82k$
$R_3 = 188.08k$
$R_4 = 203.42k$

Pins 1, 3, 17 and 18 high, pin 2 low. For 100:1 clock-to-center-frequency ratio, pin 10 is tied to V^+ and the clock frequency is 100 kHz. For 50:1 clock-to-center-frequency ratio, pin 10 is tied to V^- and the clock frequency is 50 kHz.

When building filters of order 4 or higher, best performance will always be realized when the filter blocks are cascaded in numerical order: Filter 1 (pins 16 and 15) should always precede Filter 2 (pins 5 and 6). If a second MF8 is used, Filter 2 of the first MF8 should precede Filter 1 of the second MF8, and so on.

Dynamic Considerations

Some filter response characteristics will result in high gain at certain internal nodes, particularly at the op amp output. This can cause clipping in intermediate stages even when no clipping is evident at the filter output. The consequences are significant distortion and degradation of the overall transfer function. The likelihood of clipping at the op amp output becomes greater as R_F/R_0 increases. As the design tables show, R_F/R_0 increases with increasing filter order and increasing ripple. It is good practice to keep out-of-band input signal levels small enough that the first stage can't overload.

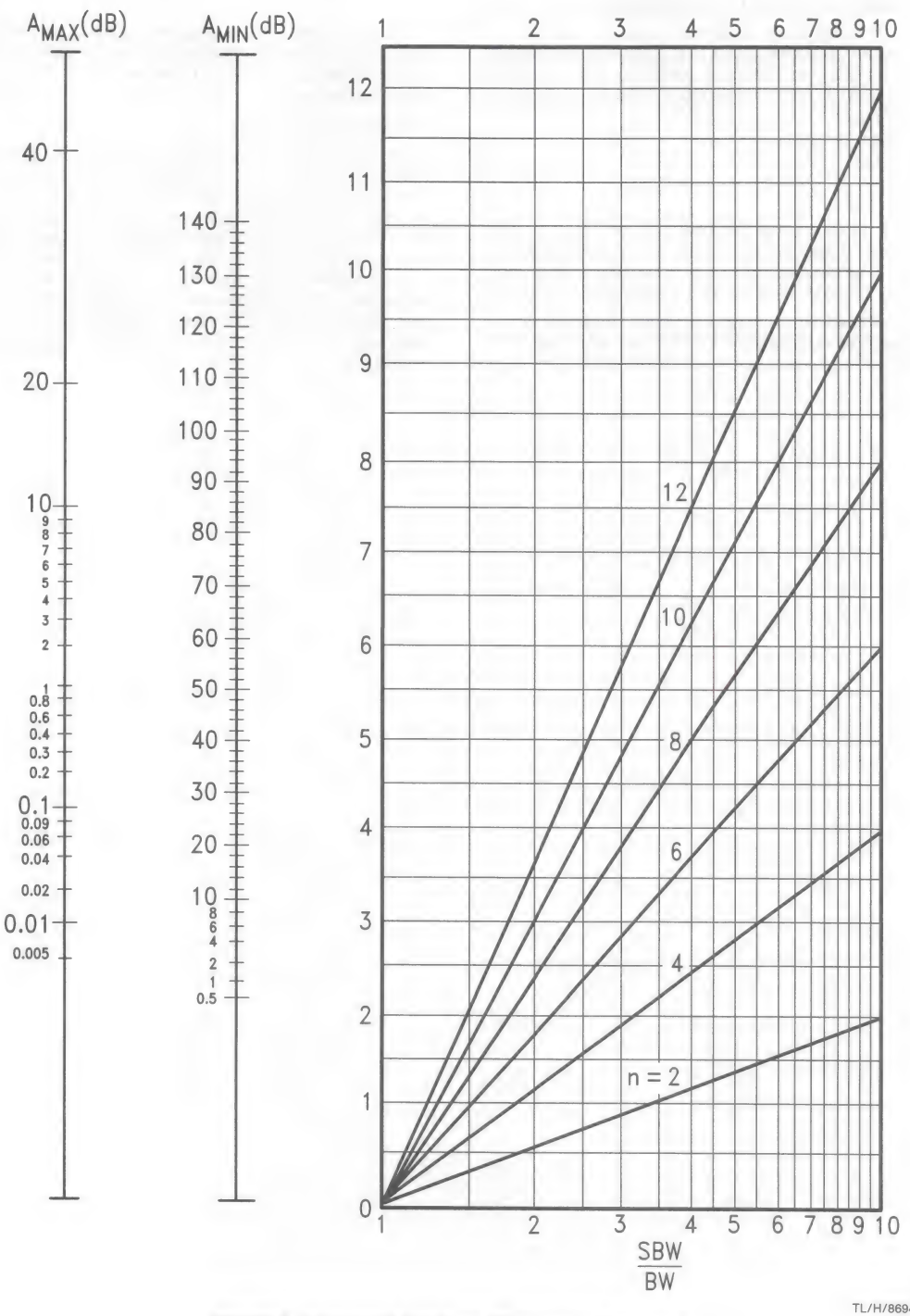


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FIGURE 8. Butterworth Bandpass Filter Design Nomograph

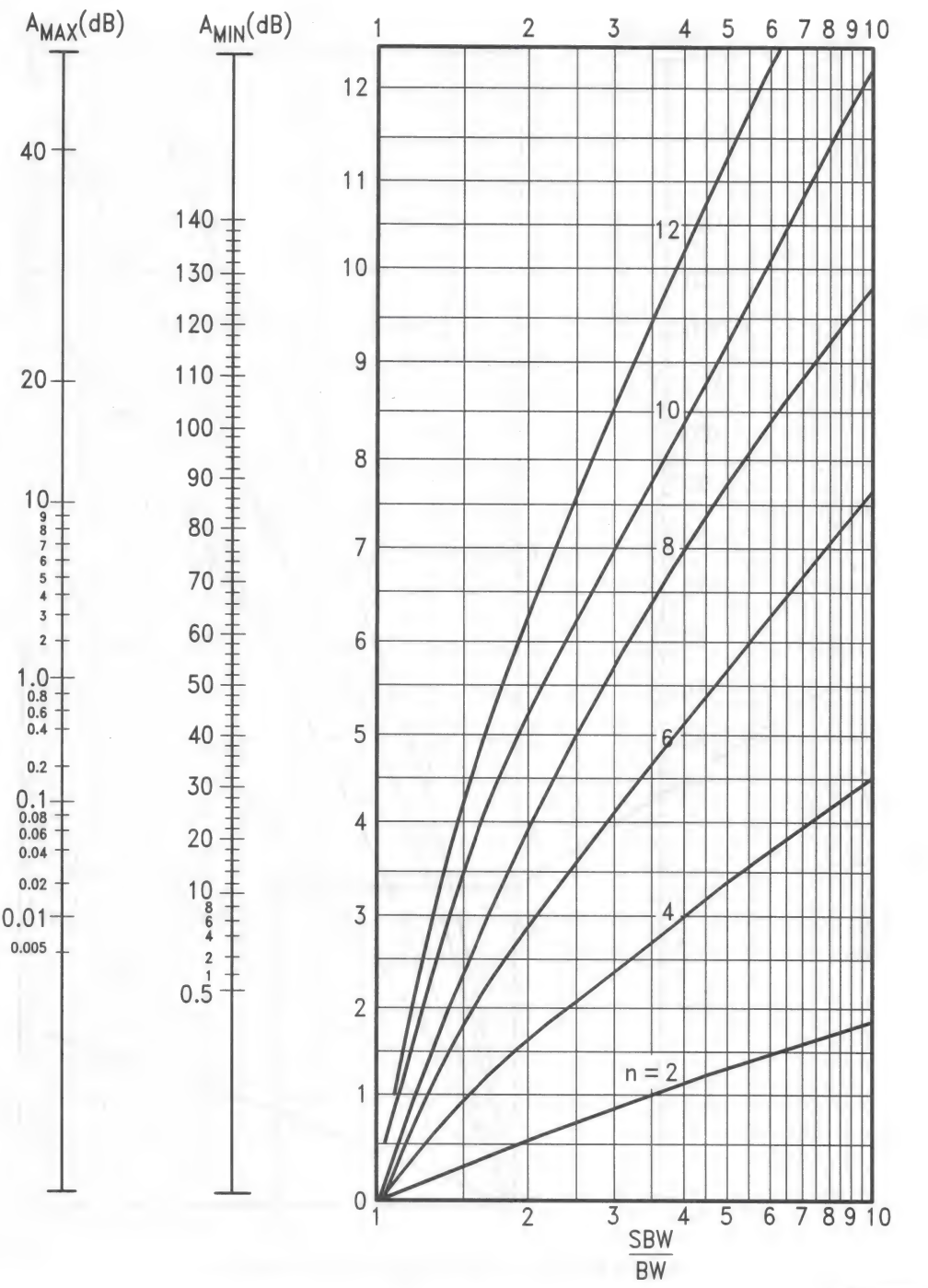

FIGURE 9. Chebyshev Bandpass Filter Design Nomograph

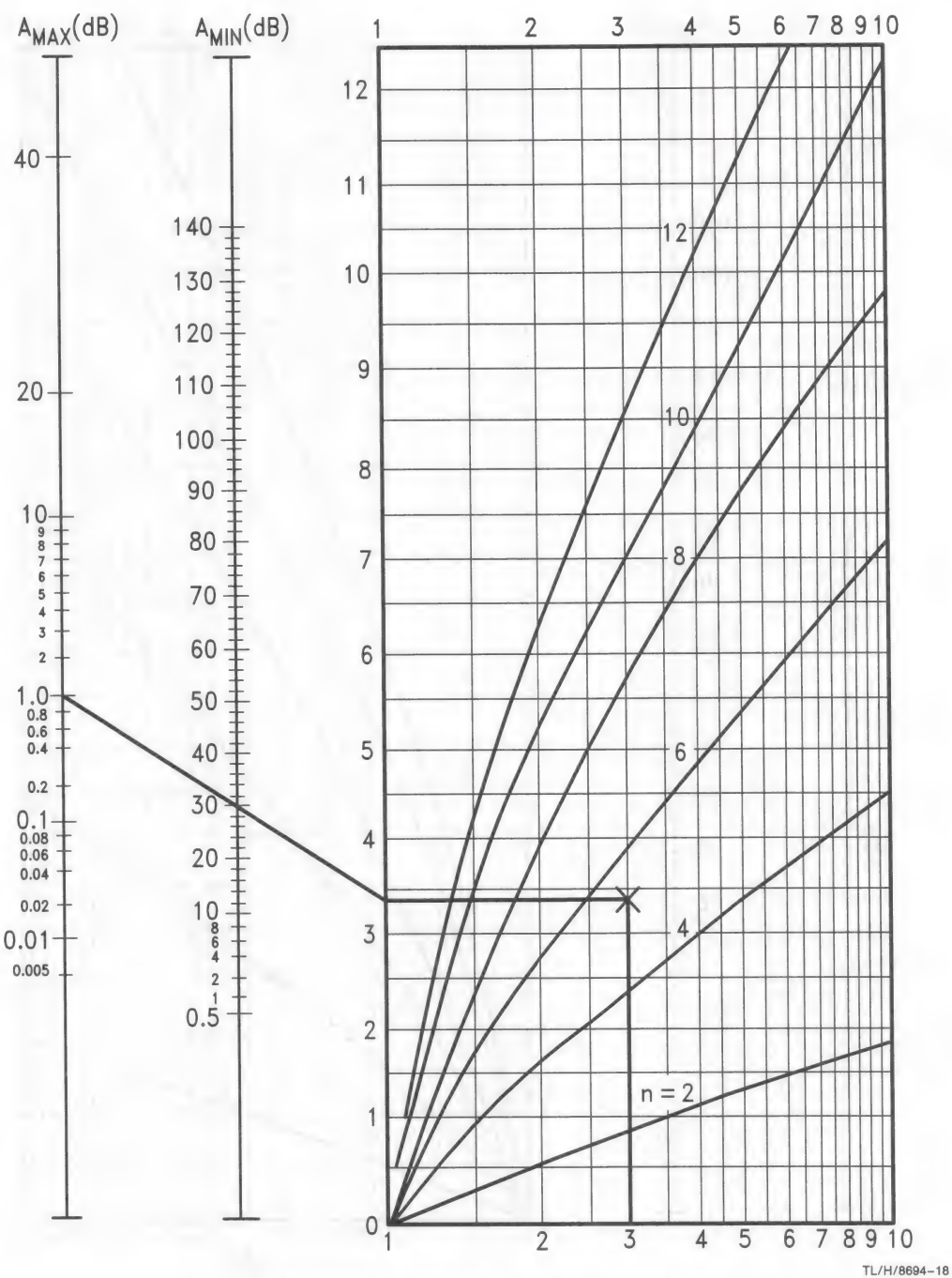

FIGURE 10. Example of Chebyshev Bandpass Nomograph Use.

A_{max} = 1 dB, A_{min} = 30 dB, and $\frac{SBW}{BW}$ = 3, resulting in n = 6.

1.0 Application Information (Continued)

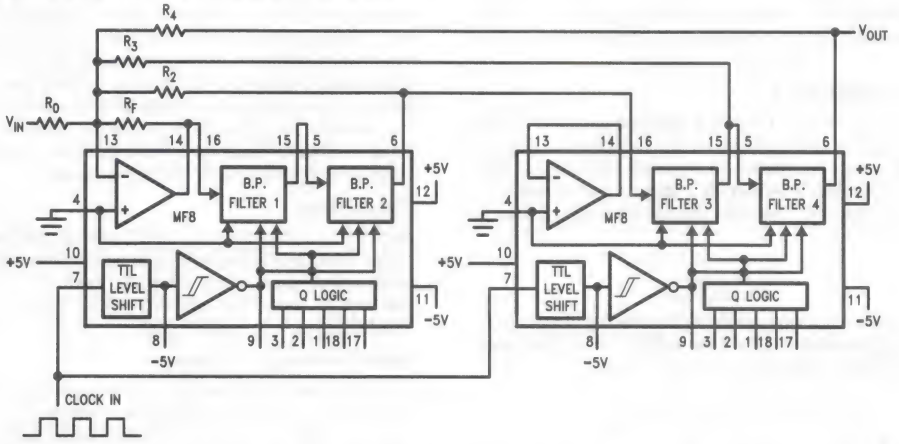

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FIGURE 11. Eighth-Order multiple-feedback bandpass filter using two MF8s. The circuit shown accepts a TTL-level clock signal and has a clock-to-center-frequency ratio of 100:1.

1.5 TRACKING AND CASCADED SECOND-ORDER BANDPASS FILTERS

The individual second-order bandpass stages may be used as "stand-alone" filters without adding external feedback resistors. The clock frequency and Q logic voltages set the center frequency and bandwidth of both second-order bandpass filters, so the two filters will have equivalent responses. Thus, they may be used as separate "tracking" filters for two different signal sources as in *Figure 2a*, or cascaded as in *Figure 2b*. For individual or cascaded second-order bandpass filters, the −3 dB bandwidth and the amplitude response are given by the following two equations:

$$BW(-3) = \frac{f_0}{Q}\sqrt{2^{(1/N)} - 1} \qquad (1)$$

$$H(s) = \left[2 \times \frac{\frac{w_0}{Q}s}{s^2 + \frac{w_0}{Q}s + w_0^2} \right]^N \qquad (2)$$

where

BW(−3) = the −3 dB bandwidth of the overall filter

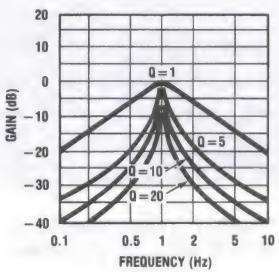

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FIGURE 12. H(s) For second-order bandpass filters with various values of Q. H_0 normalized in each case to 0 dB.

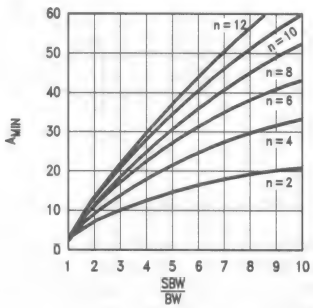

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FIGURE 13. Design Nomograph for Cascaded Identical Second-Order Bandpass Filters

Q = the Q of each second order bandpass stage

f_0 = the center frequency of the filter in Hertz

$w_0 = 2\pi f_0$ = the center frequency of the filter in radians per second

N = the number of cascaded second-order stages = $\frac{n}{2}$

H(s) = the overall filter transfer function

H(s) for a second order bandpass filter is plotted in *Figure 12*. Curves are shown for several different values of Q. Center frequency is normalized to 1 Hz and center-frequency gain is normalized to 0 dB.

To find the necessary order n for cascaded second-order bandpass filters using the nomograph in *Figure 13*, first determine the −3 dB bandwidth BW(−3), stopband width SBW, and minimum stopband attenuation A_{min}. Draw a vertical line up from SBW/BW(−3), and a horizontal line across from A_{min}. The required order is shown on the curve just above the point of intersection of the two lines. Remember that each second-order filter section will have a center frequency gain of 2, so the overall gain of a cascaded filter will be 2^N.

Cascading filters in this way may provide acceptable performance when minimum external parts count is very impor-

1.0 Application Information (Continued)

tant, but much greater flexibility and better performance will be obtained by using the feedback techniques described in 1.4.

1.6 INPUT IMPEDANCE

The input to each filter block is a switched-capacitor circuit as shown in *Figure 14*. During the first half of a clock cycle, the input capacitor charges to the input voltage V_{in}, and during the second half-cycle, its charge is transferred to a feedback capacitor. The input impedance approximates a resistor of value

$$R_{in} \cong \frac{1}{C_{in} f_{CLK}}.$$

C_{in} depends on the value of Q selected by the Q logic pins, and varies from about 1 pF to about 5 pF. For a worst-case calculation of R_{in}, assume $C_{in} = 5$ pF. Thus,

$$R_{in}(min) \cong \frac{1}{5 \times 10^{-12} f_{CLK}}$$

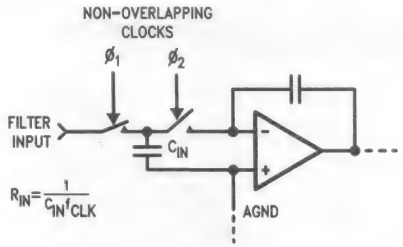

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FIGURE 14. Simplified MF8 Input Stage

At the maximum clock frequency of 1 MHz, this gives $R_{in} \cong 200$k. Note that R_{in} increases as f_{CLK} decreases, so the input impedance should never be less than this number. Source impedance should be low enough that the gain isn't significantly affected.

1.7 OUTPUT DRIVE

The filter outputs can typically drive a 5 kΩ load resistor to over ±4V peak-to-peak. Load resistors smaller than 5 kΩ should not be used. The operational amplifier can drive the minimum recommended load resistance of 5 kΩ to at least ±3.5V.

1.8 SAMPLED-DATA SYSTEM CONSIDERATIONS

Aliasing

The MF8 is a sampled-data filter, and as such, differs in many ways from conventional continuous-time filters. An important characteristic of sampled-data systems is their effect on signals at frequencies greater than one-half the sampling frequency. (The MF8's sampling frequency is the same as its clock frequency). If a signal with a frequency greater than one-half the sampling frequency is applied to the input of a sampled-data system, it will be "reflected" to a frequency less than one-half the sampling frequency. Thus, an input signal whose frequency is $f_s/2 + 10$ Hz will cause the system to respond as though the input frequency was $f_s/2 - 10$ Hz. This phenomenon is known as "aliasing". Aliasing can be reduced or eliminated by limiting the input signal spectrum to less than $f_s/2$. This may in some cases require the use of a bandwidth-limiting filter (a simple passive RC network will generally suffice) ahead of the MF8 to attenuate unwanted high-frequency signals. However, since the clock frequency is much greater than the center frequency, this will usually not be necessary.

Output Steps

Another characteristic of sampled-data circuits is that the output voltage changes only once every clock cycle, resulting in a discontinuous output signal (*Figure 15*). The "steps" are smaller when the clock-to-center-frequency ratio is 100:1 than when the ratio is 50:1.

Clock Frequency Limitations

The performance characteristics of a switched-capacitor filter depend on the switching (clock) frequency. At very low clock frequencies (below 10 Hz), the internal capacitors begin to discharge slightly between clock cycles. This is due to very small parasitic leakage currents. At very low clock frequencies, the time between clock cycles is relatively long, allowing the capacitors to discharge enough to affect the filters' output offset voltage and gain. This effect becomes stronger at elevated operating temperatures.

At higher clock frequencies, performance deviations are primarily due to the reduced time available for the internal integrating op amps to settle. For this reason, the clock waveform's duty cycle should be as close as possible to 50%, especially at higher frequencies. Filter Q shows more variation from the nominal values at higher frequencies, as indicated in the typical performance curves. This is the reason for the different maximum limits on Q accuracy at $f_{CLK} = 250$ kHz and $f_{CLK} = 100$ kHz in the table of performance specifications.

Center Frequency Accuracy

Ideally, the ratio f_{CLK}/f_0 should be precisely 100 or 50, depending on the logic voltage on pin 10. However, as Table I shows, this ratio will change slightly depending on the Q selected. As the table shows, the largest errors occur at the lowest values of Q.

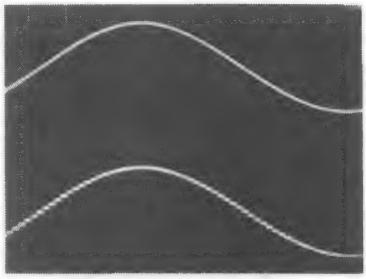

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FIGURE 15. Output Waveform of MF8 Showing Sampling Steps

2.0 Design Tables for Multiple Feedback Loop Bandpass Filters

BUTTERWORTH RIPPLE 3 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	2.0000	4.0000					1.4142
6	2.3704	2.6667	9.1429				1.5000
8	2.9142	2.0000	5.8284	14.3145			1.5307
10	3.6340	1.6000	4.4112	6.9094	27.2014		1.5451
*12	4.5635	1.3333	3.5800	4.3198	11.5043	49.0673	1.5529

CHEBYSHEV RIPPLE 0.01 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.9041	3.6339					0.4489
6	1.8277	1.8450	6.6170				0.9438
8	1.4856	0.9919	3.1209	5.0414			1.4257
*10	1.0171	0.5740	1.7484	1.2943	4.8814		1.8908

CHEBYSHEV RIPPLE 0.02 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.8644	3.4922					0.5393
6	1.7024	1.6787	6.0772				1.0849
8	1.2893	0.8707	2.7661	4.0779			1.6106
*10	0.8163	0.4934	1.5155	0.9879	3.7119		2.1179

CHEBYSHEV RIPPLE 0.03 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.8341	3.3871					0.6016
6	1.6183	1.5713	5.7231				1.1808
8	1.1688	0.7977	2.5491	3.5270			1.7362
*10	0.7034	0.4467	1.3786	0.8252	3.0938		2.2724

CHEBYSHEV RIPPLE 0.04 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.8085	3.3009					0.6508
6	1.5535	1.4908	5.4548				1.2560
8	1.0814	0.7454	2.3919	3.1471			1.8348
*10	0.6264	0.4139	1.2818	0.7181	2.6883		2.3940

CHEBYSHEV RIPPLE 0.05 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.7860	3.2268					0.6923
6	1.5002	1.4260	5.2373				1.3191
8	1.0129	0.7046	2.2685	2.8609			1.9175
*10	0.5686	0.3888	1.2072	0.6402	2.3938		2.4961

CHEBYSHEV RIPPLE 0.06 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.7657	3.1612					0.7285
6	1.4548	1.3717	5.0536				1.3741
8	0.9566	0.6713	2.1670	2.6336			1.9897
*10	0.5230	0.3685	1.1467	0.5800	2.1666		2.5852

2.0 Design Tables for Multiple Feedback Loop Bandpass Filters (Continued)

CHEBYSHEV RIPPLE .07 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.7471	3.1020					0.7609
6	1.4150	1.3249	4.8943				1.4232
8	0.9089	0.6431	2.0808	2.4466			2.0543
*10	0.4856	0.3516	1.0959	0.5316	1.9842		2.6649

CHEBYSHEV RIPPLE .08 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.7298	3.0478					0.7905
6	1.3795	1.2837	4.7534				1.4679
8	0.8675	0.6187	2.0060	2.2887			2.1130

CHEBYSHEV RIPPLE .09 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.7136	2.9978					0.8177
6	1.3475	1.2469	4.6271				1.5090
8	0.8311	0.5973	1.9400	2.1529			2.1671

CHEBYSHEV RIPPLE 0.1 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.6983	2.9512					0.8430
6	1.3183	1.2137	4.5125				1.5473
8	0.7986	0.5782	1.8809	2.0343			2.2176

CHEBYSHEV RIPPLE 0.2 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.5757	2.5998					1.0378
6	1.1128	0.9894	3.7271				1.8413
8	0.5891	0.4551	1.4954	1.3309			2.6057

CHEBYSHEV RIPPLE 0.3 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.4833	2.3575					1.1804
6	0.9835	0.8560	3.2501				2.0568
*8	0.4732	0.3861	1.2760	0.9885			2.8914

CHEBYSHEV RIPPLE 0.4 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.4067	2.1698					1.2988
6	0.8888	0.7618	2.9088				2.2363
*8	0.3956	0.3391	1.1250	0.7792			3.1299

CHEBYSHEV RIPPLE 0.5 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.3405	2.0161					1.4029
6	0.8143	0.6897	2.6447				2.3944
*8	0.3389	0.3040	1.0114	0.6365			3.3406

2.0 Design Tables for Multiple Feedback Loop Bandpass Filters (Continued)

CHEBYSHEV RIPPLE 0.6 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.2816	1.8857					1.4975
6	0.7530	0.6316	2.4305				2.5385
*8	0.2952	0.2762	0.9212	0.5326			3.5329

CHEBYSHEV RIPPLE 0.7 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.2283	1.7727					1.5852
6	0.7012	0.5834	2.2515				2.6724
*8	0.2601	0.2535	0.8471	0.4535			3.7119

CHEBYSHEV RIPPLE 0.8 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.1797	1.6731					1.6678
6	0.6564	0.5424	2.0983				2.7989
*8	0.2314	0.2344	0.7846	0.3913			3.8811

CHEBYSHEV RIPPLE 0.9 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.1347	1.5841					1.7464
6	0.6171	0.5068	1.9650				2.9194
*8	0.2073	0.2181	0.7309	0.3413			4.0426

CHEBYSHEV RIPPLE 1.0 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.0930	1.5039					1.8219
6	0.5822	0.4756	1.8475				3.0354
*8	0.1869	0.2038	0.6840	0.3002			4.1981

CHEBYSHEV RIPPLE 1.1 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.0539	1.4310					1.8949
6	0.5509	0.4479	1.7428				3.1476
*8	0.1693	0.1913	0.6426	0.2660			4.3487

CHEBYSHEV RIPPLE 1.2 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	1.0173	1.3643					1.9657
6	0.5226	0.4231	1.6487				3.2567
*8	0.1540	0.1801	0.6056	0.2372			4.4952

CHEBYSHEV RIPPLE 1.3 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.9828	1.3029					2.0348
6	0.4969	0.4006	1.5634				3.3633
*8	0.1406	0.1701	0.5724	0.2125			4.6385

2.0 Design Tables for Multiple Feedback Loop Bandpass Filters (Continued)

CHEBYSHEV RIPPLE 1.4 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.9501	1.2461					2.1024
6	0.4733	0.3803	1.4857				3.4678

CHEBYSHEV RIPPLE 1.5 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.9192	1.1934					2.1688
6	0.4515	0.3616	1.4145				3.5705

CHEBYSHEV RIPPLE 1.6 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.8897	1.1443					2.2341
6	0.4315	0.3445	1.3490				3.6717

CHEBYSHEV RIPPLE 1.7 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.8617	1.0983					2.2986
6	0.4128	0.3287	1.2883				3.7717

CHEBYSHEV RIPPLE 1.8 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.8350	1.0553					2.3624
6	0.3955	0.3141	1.2321				3.8706

CHEBYSHEV RIPPLE 1.9 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.8095	1.0148					2.4255
6	0.3793	0.3005	1.1797				3.9687

CHEBYSHEV RIPPLE 2.0 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.7850	0.9767					2.4881
6	0.3641	0.2878	1.1308				4.0660

CHEBYSHEV RIPPLE 2.1 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.7616	0.9407					2.5503
6	0.3498	0.2759	1.0850				4.1628

CHEBYSHEV RIPPLE 2.2 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.7391	0.9067					2.6122
6	0.3364	0.2648	1.0420				4.2591

2.0 Design Tables for Multiple Feedback Loop Bandpass Filters (Continued)

CHEBYSHEV RIPPLE 2.3 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.7176	0.8744					2.6737
6	0.3237	0.2544	1.0016				4.3550

CHEBYSHEV RIPPLE 2.4 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.6968	0.8438					2.7350
6	0.3118	0.2446	0.9635				4.4507

CHEBYSHEV RIPPLE 2.5 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.6769	0.8148					2.7962
6	0.3005	0.2353	0.9275				4.5462

CHEBYSHEV RIPPLE 2.6 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.6577	0.7871					2.8573
6	0.2897	0.2265	0.8935				4.6415

CHEBYSHEV RIPPLE 2.7 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.6392	0.7607					2.9183
6	0.2796	0.2182	0.8612				4.7368

CHEBYSHEV RIPPLE 2.8 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.6213	0.7356					2.9792
6	0.2699	0.2104	0.8306				4.8322

CHEBYSHEV RIPPLE 2.9 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.6041	0.7116					3.0402
6	0.2607	0.2029	0.8016				4.9276

CHEBYSHEV RIPPLE 3.0 dB

Order	K_0	K_2	K_3	K_4	K_5	K_6	K_Q
4	0.5875	0.6886					3.1013
6	0.2519	0.1959	0.7739				5.0231

Note: Multiple feedback loop filters of higher order than those specified in the tables will oscillate due to phase shift at the output of the summing amplifier. This phase shift is not the fault of the MF8; it is inherent in this type of multiple feedback loop topology. In addition, all filters marked with an asterisk (*) will be unstable for $Q \leq 1$, due to phase shifts caused by the MF8's switched-capacitor design approach.

National Semiconductor

MF10
Universal Monolithic Dual Switched Capacitor Filter

General Description

The MF10 consists of 2 independent and extremely easy to use, general purpose CMOS active filter building blocks. Each block, together with an external clock and 3 to 4 resistors, can produce various 2nd order functions. Each building block has 3 output pins. One of the outputs can be configured to perform either an allpass, highpass or a notch function; the remaining 2 output pins perform lowpass and bandpass functions. The center frequency of the lowpass and bandpass 2nd order functions can be either directly dependent on the clock frequency, or they can depend on both clock frequency and external resistor ratios. The center frequency of the notch and allpass functions is directly dependent on the clock frequency, while the highpass center frequency depends on both resistor ratio and clock. Up to 4th order functions can be performed by cascading the two 2nd order building blocks of the MF10; higher than 4th order functions can be obtained by cascading MF10 packages.

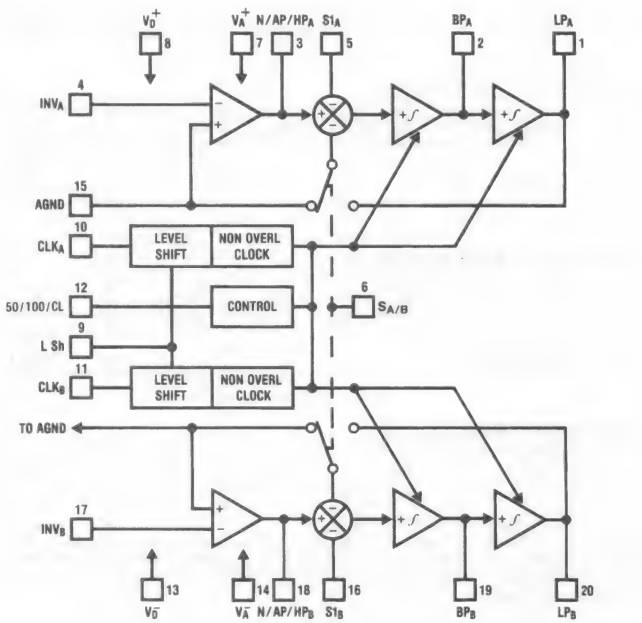
Any of the classical filter configurations (such as Butterworth, Bessel, Cauer and Chebyshev) can be formed.

For pin-compatible device with improved performance refer to LMF100 datasheet.

Features

- Easy to use
- Clock to center frequency ratio accuracy ±0.6%
- Filter cutoff frequency stability directly dependent on external clock quality
- Low sensitivity to external component variation
- Separate highpass (or notch or allpass), bandpass, lowpass outputs
- $f_O \times Q$ range up to 200 kHz
- Operation up to 30 kHz
- 20-pin 0.3″ wide Dual-In-Line package
- 20-pin Surface Mount (SO) wide-body package

System Block Diagram



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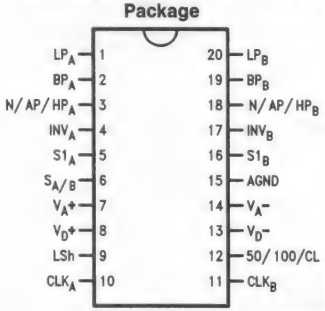
Connection Diagram

Surface Mount and Dual-In-Line Package



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Top View

Order Number MF10AJ or MF10CCJ
See NS Package Number J20A

Order Number MF10CCWM
See NS Package Number M20B

Order Number MF10ACN or MF10CCN
See NS Package Number N20A

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage ($V^+ - V^-$)	14V
Voltage at Any Pin	$V^+ + 0.3V$ $V^- - 0.3V$
Input Current at Any Pin (Note 2)	5 mA
Package Input Current (Note 2)	20 mA
Power Dissipation (Note 3)	500 mW
Storage Temperature	150°C
ESD Susceptability (Note 11)	2000V

Soldering Information

N Package: 10 sec.	260°C
J Package: 10 sec.	300°C
SO Package: Vapor Phase (60 Sec.)	215°C
Infrared (15 Sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (Appendix D) for other methods of soldering surface mount devices.

Operating Ratings (Note 1)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
MF10ACN, MF10CCN	$0°C \le T_A \le 70°C$
MF10CCWM	$0°C \le T_A \le 70°C$
MF10CCJ	$-40°C \le T_A \le 85°C$
MF10AJ	$-55°C \le T_A \le 125°C$

Electrical Characteristics

$V^+ = +5.00V$ and $V^- = -5.00V$ unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Symbol	Parameter		Conditions		MF10ACN, MF10CCN, MF10CCWM			MF10CCJ, MF10AJ			Units
					Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
$V^+ - V^-$	Supply Voltage	Min					**8**			**8**	V
		Max					**14**			**14**	V
I_S	Maximum Supply Current		Clock Applied to Pins 10 & 11 No Input Signal		8	12	**12**	8	**12**		mA
f_O	Center Frequency Range	Min	$f_O \times Q < 200$ kHz		0.1		**0.2**	0.1		**0.2**	Hz
		Max			30		**20**	30		**20**	kHz
f_{CLK}	Clock Frequency Range	Min			5.0		**10**	5.0		**10**	Hz
		Max			1.5		**1.0**	1.5		**1.0**	MHz
f_{CLK}/f_O	50:1 Clock to Center Frequency Ratio Deviation	MF10A	Q = 10 Mode 1	$V_{pin12} = 5V$ $f_{CLK} = 250$ kHz	±0.2	±0.6	**±0.6**	±0.2	**±1.0**		%
		MF10C			±0.2	±1.5	**±1.5**	±0.2	**±1.5**		%
f_{CLK}/f_O	100:1 Clock to Center Frequency Ratio Deviation	MF10A	Q = 10 Mode 1	$V_{pin12} = 0V$ $f_{CLK} = 500$ kHz	±0.2	±0.6	**±0.6**	±0.2	**±1.0**		%
		MF10C			±0.2	±1.5	**±1.5**	±0.2	**±1.5**		%
	Clock Feedthrough		Q = 10 Mode 1		10			10			mV
	Q Error (MAX) (Note 4)		Q = 10 Mode 1	$V_{pin12} = 5V$ $f_{CLK} = 250$ kHz	±2	±6	**±6**	±2	**±6**		%
				$V_{pin12} = 0V$ $f_{CLK} = 500$ kHz	±2	±6	**±6**	±2	**±6**		%
H_{OLP}	DC Lowpass Gain		Mode 1 R1 = R2 = 10k		0	±0.2	**±0.2**	0	**±0.2**		dB
V_{OS1}	DC Offset Voltage (Note 5)				±5.0	±15	**±15**	±5.0	**±15**		mV
V_{OS2}	DC Offset Voltage (Note 5)	Min	$V_{pin12} = +5V$ $(f_{CLK}/f_O = 50)$	$S_{A/B} = V^+$	−150	−185	**−185**	−150	**−185**		mV
		Max				−85	**−85**		**−85**		
		Min	$V_{pin12} = +5V$ $(f_{CLK}/f_O = 50)$	$S_{A/B} = V^-$	−70			−70			mV
		Max									
V_{OS3}	DC Offset Voltage (Note 5)	Min	$V_{pin12} = +5V$ $(f_{CLK}/f_O = 50)$	All Modes	−70	−100	**−100**	−70	**−100**		mV
		Max				−20	**−20**		**−20**		
V_{OS2}	DC Offset Voltage (Note 5)		$V_{pin12} = 0V$ $(f_{CLK}/f_O = 100)$	$S_{A/B} = V^+$	−300			−300			mV
	(Note 5)		$V_{pin12} = 0V$ $(f_{CLK}/f_O = 100)$	$S_{A/B} = V^-$	−140			−140			mV
V_{OS3}	DC Offset Voltage (Note 5)		$V_{pin12} = 0V$ $(f_{CLK}/f_O = 100)$	All Modes	−140			−140			mV

Electrical Characteristics (Continued) $V^+ = +5.00V$ and $V^- = -5.00V$ unless otherwise specified.

Boldface limits apply for T_{MIN} to T_{MAX}; all other limits $T_A = T_J = 25°C$.

Symbol	Parameter		Conditions	MF10ACN, MF10CCN, MF10CCWM			MF10CCJ, MF10AJ			Units
				Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
V_{OUT}	Minimum Output Voltage Swing	BP, LP Pins	$R_L = 5k$	±4.25	±3.8	**±3.8**	±4.25	**±3.8**		V
		N/AP/HP Pin	$R_L = 3.5k$	±4.25	±3.8	**±3.8**	±4.25	**±3.6**		V
GBW	Op Amp Gain BW Product			2.5			2.5			MHz
SR	Op Amp Slew Rate			7			7			V/μs
	Dynamic Range (Note 6)		$V_{pin12} = +5V$ ($f_{CLK}/f_O = 50$)	83			83			dB
			$V_{pin12} = 0V$ ($f_{CLK}/f_O = 100$)	80			80			dB
I_{SC}	Maximum Output Short Circuit Current (Note 7)	Source		20			20			mA
		Sink		3.0			3.0			mA

Logic Input Characteristics **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$

Parameter		Conditions	MF10ACN, MF10CCN, MF10CCWM			MF10CCJ, MF10AJ			Units
			Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typical (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
CMOS Clock Input Voltage	Min Logical "1"	$V^+ = +5V, V^- = -5V,$ $V_{LSh} = 0V$		+3.0	**+3.0**		**+3.0**		V
	Max Logical "0"			−3.0	**−3.0**		**−3.0**		V
	Min Logical "1"	$V^+ = +10V, V^- = 0V,$ $V_{LSh} = +5V$		+8.0	**+8.0**		**+8.0**		V
	Max Logical "0"			+2.0	**+2.0**		**+2.0**		V
TTL Clock Input Voltage	Min Logical "1"	$V^+ = +5V, V^- = -5V,$ $V_{LSh} = 0V$		+2.0	**+2.0**		**+2.0**		V
	Max Logical "0"			+0.8	**+0.8**		**+0.8**		V
	Min Logical "1"	$V^+ = +10V, V^- = 0V,$ V_{LSh}		+2.0	**+2.0**		**+2.0**		V
	Max Logical "0"			+0.8	**+0.8**		**+0.8**		V

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 5 mA or less. The 20 mA package input current limits the number of pins that can exceed the power supply boundaries with a 5 mA current limit to four.

Note 3: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 125°C$, and the typical junction-to-ambient thermal resistance of the MF10ACN/CCN when board mounted is 55°C/W. For the MF10AJ/CCJ, this number increases to 95°C/W and for the MF10CCWM this number is 66°C/W.

Note 4: The accuracy of the Q value is a function of the center frequency (f_O). This is illustrated in the curves under the heading "Typical Performance Characteristics".

Note 5: V_{OS1}, V_{OS2}, and V_{OS3} refer to the internal offsets as discussed in the Applications Information Section 3.4.

Note 6: For ±5V supplies the dynamic range is referenced to 2.82V rms (4V peak) where the wideband noise over a 20 kHz bandwidth is typically 200 μV rms for the MF10 with a 50:1 CLK ratio and 280 μV rms for the MF10 with a 100:1 CLK ratio.

Note 7: The short circuit source current is measured by forcing the output that is being tested to its maximum positive voltage swing and then shorting that output to the negative supply. The short circuit sink current is measured by forcing the output that is being tested to its maximum negative voltage swing and then shorting that output to the positive supply. These are the worst case conditions.

Note 8: Typicals are at 25°C and represent most likely parametric norm.

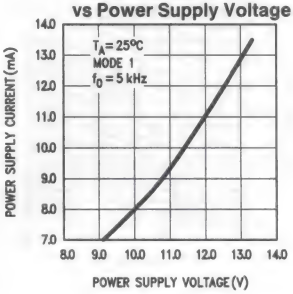
Note 9: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 10: Design limits are guaranteed but not 100% tested. These limits are not used to calculate outgoing quality levels.

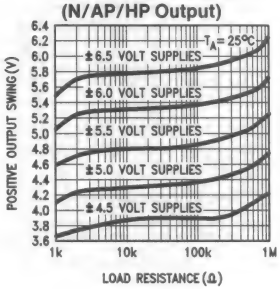
Note 11: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Typical Performance Characteristics

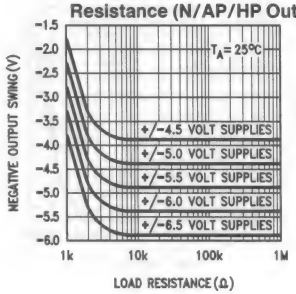
Power Supply Current vs Power Supply Voltage



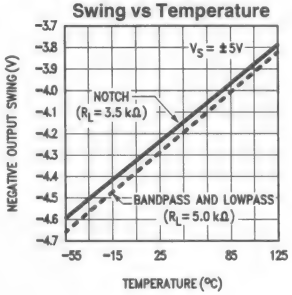
Positive Output Voltage Swing vs Load Resistance (N/AP/HP Output)



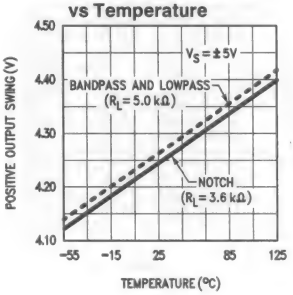
Negative Output Voltage Swing vs Load Resistance (N/AP/HP Output)



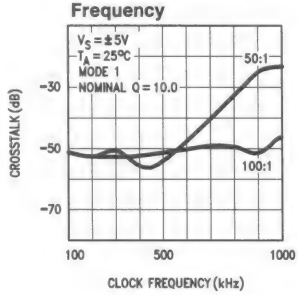
Negative Output Swing vs Temperature



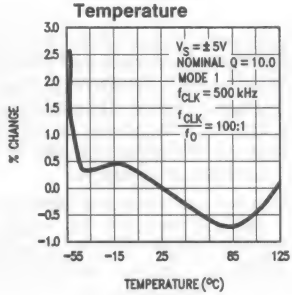
Positive Output Swing vs Temperature



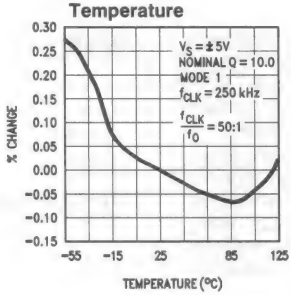
Crosstalk vs Clock Frequency



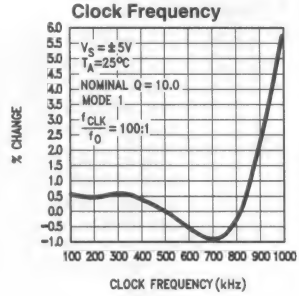
Q Deviation vs Temperature



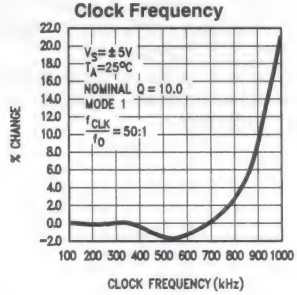
Q Deviation vs Temperature



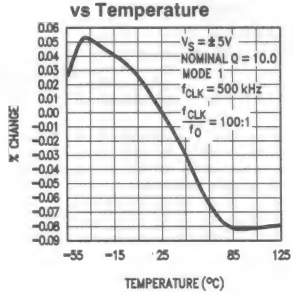
Q Deviation vs Clock Frequency



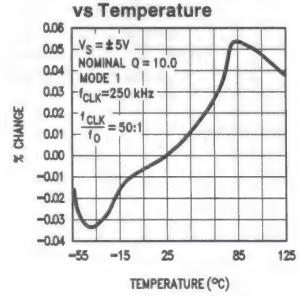
Q Deviation vs Clock Frequency



f_{CLK}/f_O Deviation vs Temperature



f_{CLK}/f_O Deviation vs Temperature



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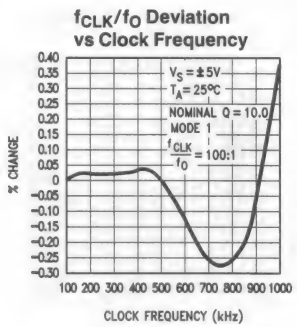
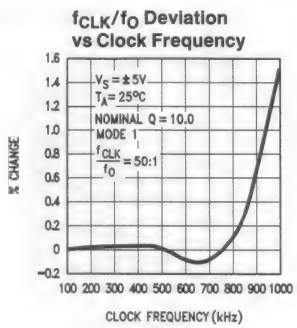
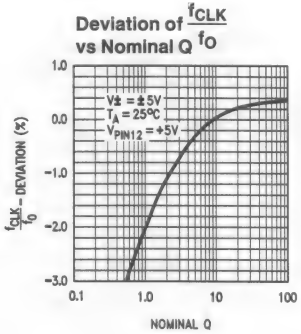
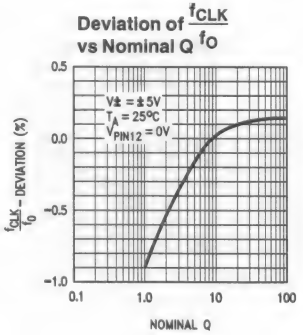
Typical Performance Characteristics (Continued)









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Pin Descriptions

LP(1,20), BP(2,19), N/AP/HP(3,18) — The second order lowpass, bandpass and notch/allpass/highpass outputs. These outputs can typically sink 1.5 mA and source 3 mA. Each output typically swings to within 1V of each supply.

INV(4,17) — The inverting input of the summing op-amp of each filter. These are high impedance inputs, but the non-inverting input is internally tied to AGND, making INV_A and INV_B behave like summing junctions (low impedance, current inputs).

S1(5,16) — S1 is a signal input pin used in the all-pass filter configurations (see modes 4 and 5). The pin should be driven with a source impedance of less than 1 kΩ. If S1 is not driven with a signal it should be tied to AGND (mid-supply).

$S_{A/B}$(6) — This pin activates a switch that connects one of the inputs of each filter's second summer to either AGND ($S_{A/B}$ tied to V^-) or to the lowpass (LP) output ($S_{A/B}$ tied to V^+). This offers the flexibility needed for configuring the filter in its various modes of operation.

V_A^+(7), V_D^+(8) — Analog positive supply and digital positive supply. These pins are internally connected through the IC substrate and therefore V_A^+ and V_D^+ should be derived from the same power supply source. They have been brought out separately so they can be bypassed by separate capacitors, if desired. They can be externally tied together and bypassed by a single capacitor.

V_A^-(14), V_D^-(13) — Analog and digital negative supplies. The same comments as for V_A^+ and V_D^+ apply here.

Pin Descriptions (Continued)

LSh(9) — Level shift pin; it accommodates various clock levels with dual or single supply operation. With dual ±5V supplies, the MF10 can be driven with CMOS clock levels (±5V) and the LSh pin should be tied to the system ground. If the same supplies as above are used but only TTL clock levels, derived from 0V to +5V supply, are available, the LSh pin should be tied to the system ground. For single supply operation (0V and +10V) the V_A^-, V_D^- pins should be connected to the system ground, the AGND pin should be biased at +5V and the LSh pin should also be tied to the system ground for TTL clock levels. LSh should be biased at +5V for CMOS clock levels in 10V single-supply applications.

CLKA(10), CLKB(11) — Clock inputs for each switched capacitor filter building block. They should both be of the same level (TTL or CMOS). The level shift (LSh) pin description discusses how to accommodate their levels. The duty cycle of the clock should be close to 50% especially when clock frequencies above 200 kHz are used. This allows the maximum time for the internal op-amps to settle, which yields optimum filter operation.

50/100/CL(12) — By tying this pin high a 50:1 clock-to-filter-center-frequency ratio is obtained. Tying this pin at mid-supplies (i.e, analog ground with dual supplies) allows the filter to operate at a 100:1 clock-to-center-frequency ratio. When the pin is tied low (i.e., negative supply with dual supplies), a simple current limiting circuit is triggered to limit the overall supply current down to about 2.5 mA. The filtering action is then aborted.

AGND(15) — This is the analog ground pin. This pin should be connected to the system ground for dual supply operation or biased to mid-supply for single supply operation. For a further discussion of mid-supply biasing techniques see the Applications Information (Section 3.2). For optimum filter performance a "clean" ground must be provided.

1.0 Definition of Terms

f_{CLK}: the frequency of the external clock signal applied to pin 10 or 11.

f_O: center frequency of the second order function complex pole pair. f_O is measured at the bandpass outputs of the MF10, and is the frequency of maximum bandpass gain. *(Figure 1)*

f_{notch}: the frequency of minimum (ideally zero) gain at the notch outputs.

f_Z: the center frequency of the second order complex zero pair, if any. If f_Z is different from f_O and if Q_Z is high, it can be observed as the frequency of a notch at the allpass output. *(Figure 10)*

Q: "quality factor" of the 2nd order filter. Q is measured at the bandpass outputs of the MF10 and is equal to f_O divided by the −3 dB bandwidth of the 2nd order bandpass filter *(Figure 1)*. The value of Q determines the shape of the 2nd order filter responses as shown in *Figure 6*.

Q_Z: the quality factor of the second order complex zero pair, if any. Q_Z is related to the allpass characteristic, which is written:

$$H_{AP}(s) = \frac{H_{OAP}\left(s^2 - \frac{s\omega_O}{Q_Z} + \omega_O^2\right)}{s^2 + \frac{s\omega_O}{Q} + \omega_O^2}$$

where $Q_Z = Q$ for an all-pass response.

H_{OBP}: the gain (in V/V) of the bandpass output at $f = f_O$.

H_{OLP}: the gain (in V/V) of the lowpass output as $f \rightarrow 0$ Hz *(Figure 2)*.

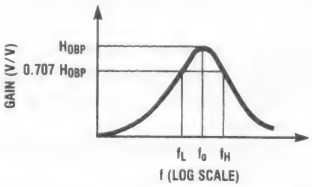
H_{OHP}: the gain (in V/V) of the highpass output as $f \rightarrow f_{CLK}/2$ *(Figure 3)*.

H_{ON}: the gain (in V/V) of the notch output as $f \rightarrow 0$ Hz and as $f \rightarrow f_{CLK}/2$, when the notch filter has equal gain above and below the center frequency *(Figure 4)*. When the low-frequency gain differs from the high-frequency gain, as in modes 2 and 3a *(Figures 11 and 8)*, the two quantities below are used in place of H_{ON}.

H_{ON1}: the gain (in V/V) of the notch output as $f \rightarrow 0$ Hz.

H_{ON2}: the gain (in V/V) of the notch output as $f \rightarrow f_{CLK}/2$.

1.0 Definition of Terms (Continued)



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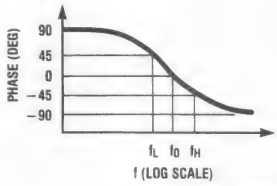
(a)



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(b)

$$H_{BP}(s) = \frac{H_{OBP}s}{s^2 + \frac{s\omega_O}{Q} + \omega_O^2}$$

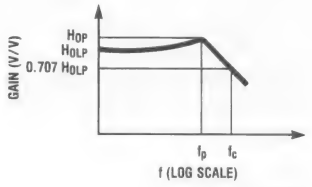
$$Q = \frac{f_O}{f_H - f_L};\ f_O = \sqrt{f_L f_H}$$

$$f_L = f_O\left(\frac{-1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

$$f_H = f_O\left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

$$\omega_O = 2\pi f_O$$

FIGURE 1. 2nd-Order Bandpass Response



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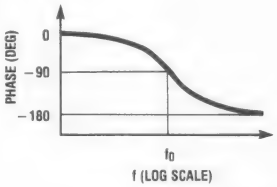
(a)



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(b)

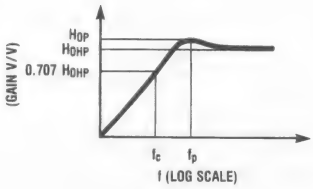
$$H_{LP}(s) = \frac{H_{OLP}\omega_O^2}{s^2 + \frac{s\omega_O}{Q} + \omega_O^2}$$

$$f_C = f_O \times \sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}$$

$$f_p = f_O\sqrt{1 - \frac{1}{2Q^2}}$$

$$H_{OP} = H_{OLP} \times \frac{1}{\frac{1}{Q}\sqrt{1 - \frac{1}{4Q^2}}}$$

FIGURE 2. 2nd-Order Low-Pass Response



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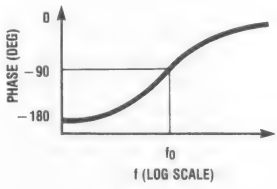
(a)



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(b)

$$H_{HP}(s) = \frac{H_{OHP}s^2}{s^2 + \frac{s\omega_O}{Q} + \omega_O^2}$$

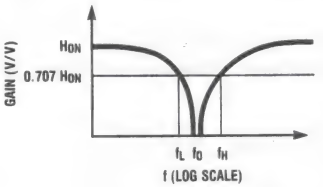
$$f_C = f_O \times \left[\sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}\right]^{-1}$$

$$f_p = f_O \times \left[\sqrt{1 - \frac{1}{2Q^2}}\right]^{-1}$$

$$H_{OP} = H_{OHP} \times \frac{1}{\frac{1}{Q}\sqrt{1 - \frac{1}{4Q^2}}}$$

FIGURE 3. 2nd-Order High-Pass Response

1.0 Definitions of Terms (Continued)



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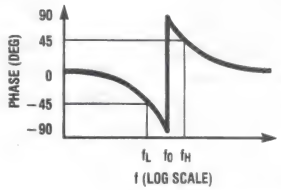
(a)



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(b)

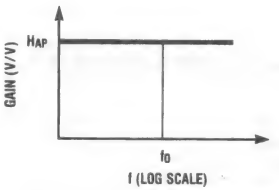
$$H_N(s) = \frac{H_{ON}(s^2 + \omega_O^2)}{s^2 + \frac{s\omega_O}{Q} + \omega_O^2}$$

$$Q = \frac{f_O}{f_H - f_L};\ f_O = \sqrt{f_L f_H}$$

$$f_L = f_O\left(\frac{-1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

$$f_H = f_O\left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1}\right)$$

FIGURE 4. 2nd-Order Notch Response



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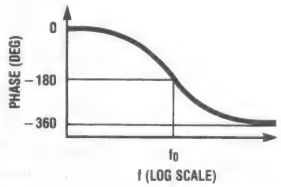
(a)



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(b)

$$H_{AP}(s) = \frac{H_{OAP}\left(s^2 - \frac{s\omega_O}{Q} + \omega_O^2\right)}{s^2 + \frac{s\omega_O}{Q} + \omega_O^2}$$

FIGURE 5. 2nd-Order All-Pass Response

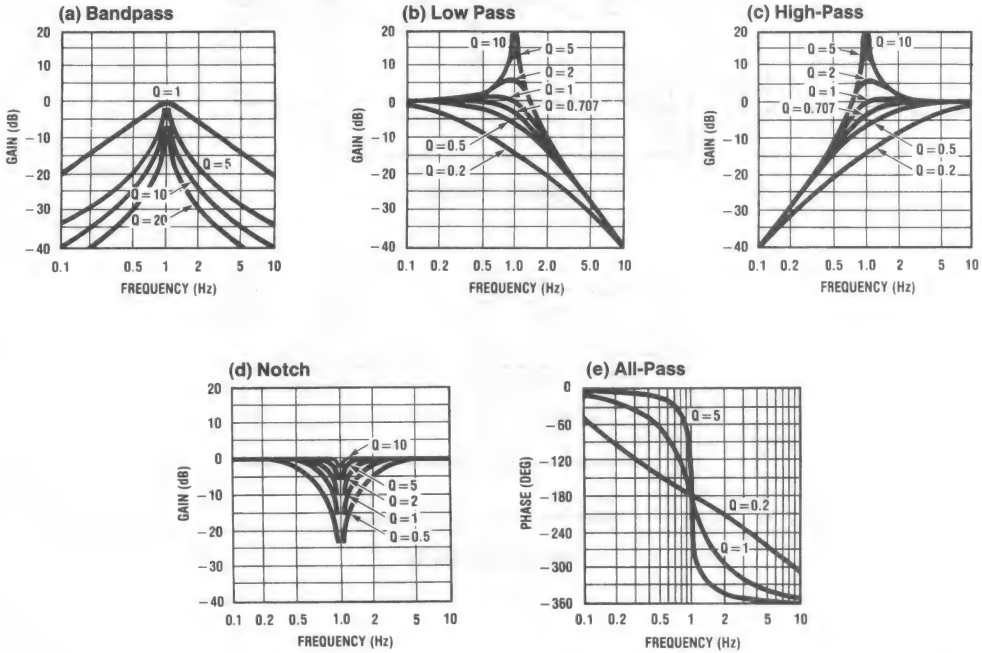

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FIGURE 6. Response of various 2nd-order filters as a function of Q. Gains and center frequencies are normalized to unity.

2.0 Modes of Operation

The MF10 is a switched capacitor (sampled data) filter. To fully describe its transfer functions, a time domain approach is appropriate. Since this is cumbersome, and since the MF10 closely approximates continuous filters, the following discussion is based on the well know frequency domain. Each MF10 can produce a full 2nd order function. See Table I for a summary of the characteristics of the various modes.

MODE 1: Notch 1, Bandpass, Lowpass Outputs: $f_{notch} = f_O$ (See *Figure 7*)

f_O = center frequency of the complex pole pair

$= \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

f_{notch} = center frequency of the imaginary zero pair = f_O.

H_{OLP} = Lowpass gain (as $f \rightarrow 0$) $= -\frac{R2}{R1}$

H_{OBP} = Bandpass gain (at $f = f_O$) $= -\frac{R3}{R1}$

H_{ON} = Notch output gain as $\left.\begin{matrix} f \rightarrow 0 \\ f \rightarrow f_{CLK}/2 \end{matrix}\right\} = \frac{-R_2}{R_1}$

$Q = \frac{f_O}{BW} = \frac{R3}{R2}$

= quality factor of the complex pole pair

BW = the −3 dB bandwidth of the bandpass output.

Circuit dynamics:

$H_{OLP} = \frac{H_{OBP}}{Q}$ or $H_{OBP} = H_{OLP} \times Q$

$= H_{ON} \times Q.$

$H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

MODE 1a: Non-Inverting BP, LP (See *Figure 8*)

$f_O = \frac{f_{CLK}}{100}$ or $\frac{f_{CLK}}{50}$

$Q = \frac{R3}{R2}$

$H_{OLP} = -1$; $H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

$H_{OBP_1} = -\frac{R3}{R2}$

$H_{OBP_2} = 1$ (Non-Inverting)

Circuit Dynamics: $H_{OBP1} = Q$

Note: V_{IN} should be driven from a low impedance (<1 kΩ) source.

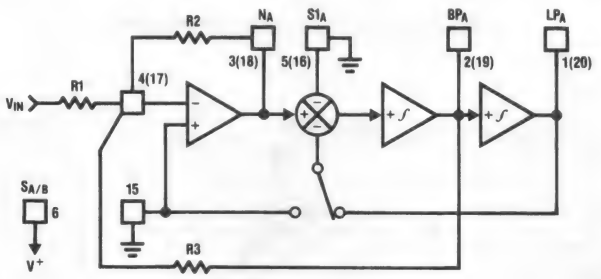

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FIGURE 7. MODE 1

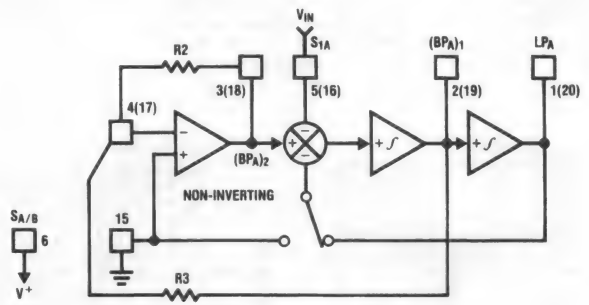

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FIGURE 8. MODE 1a

2.0 Modes of Operation (Continued)

MODE 2: Notch 2, Bandpass, Lowpass: $f_{notch} < f_O$ (See *Figure 9*)

f_O = center frequency

$$= \frac{f_{CLK}}{100}\sqrt{\frac{R2}{R4}+1} \text{ or } \frac{f_{CLK}}{50}\sqrt{\frac{R2}{R4}+1}$$

$$f_{notch} = \frac{f_{CLK}}{100} \text{ or } \frac{f_{CLK}}{50}$$

Q = quality factor of the complex pole pair

$$= \frac{\sqrt{R2/R4+1}}{R2/R3}$$

H_{OLP} = Lowpass output gain (as $f \rightarrow 0$)

$$= -\frac{R2/R1}{R2/R4+1}$$

H_{OBP} = Bandpass output gain (at $f = f_O$) $= -R3/R1$

H_{ON_1} = Notch output gain (as $f \rightarrow 0$)

$$= -\frac{R2/R1}{R2/R4+1}$$

H_{ON_2} = Notch output gain $\left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right) = -R2/R1$

Filter dynamics: $H_{OBP} = Q\sqrt{H_{OLP}\,H_{ON_2}} = \sqrt{H_{ON1}\,H_{ON2}}$

MODE 3: Highpass, Bandpass, Lowpass Outputs (See *Figure 10*)

$$f_O = \frac{f_{CLK}}{100}\times\sqrt{\frac{R2}{R4}} \text{ or } \frac{f_{CLK}}{50}\times\sqrt{\frac{R2}{R4}}$$

Q = quality factor of the complex pole pair

$$= \sqrt{\frac{R2}{R4}}\times\frac{R3}{R2}$$

$$H_{OHP} = \text{Highpass Gain}\left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right) = -\frac{R2}{R1}$$

$$H_{OBP} = \text{Lowpass Gain}\left(\text{at } f = f_O\right) = -\frac{R3}{R1}$$

$$H_{OLP} = \text{Lowpass Gain}\left(\text{as } f \rightarrow 0\right) = -\frac{R4}{R1}$$

Circuit dynamics: $\frac{R2}{R4} = \frac{H_{OHP}}{H_{OLP}}$;

$$H_{OBP} = \sqrt{H_{OHP}\times H_{OLP}}\times Q$$

$H_{OLP(peak)} \cong Q \times H_{OLP}$ (for high Q's)

$H_{OHP(peak)} \cong Q \times H_{OHP}$ (for high Q's)

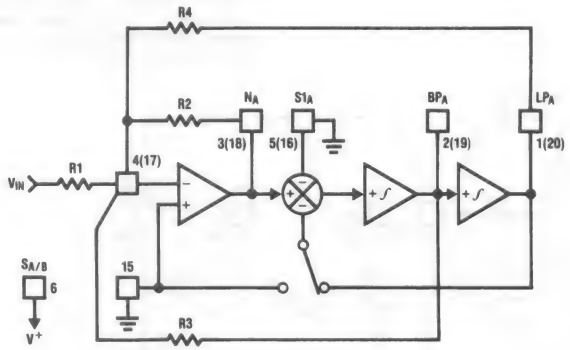

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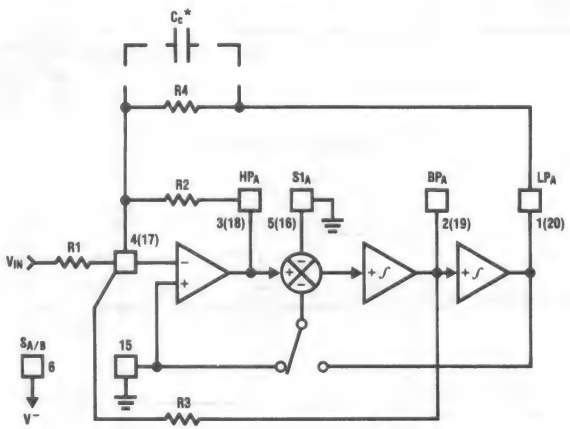
FIGURE 9. MODE 2



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*In Mode 3, the feedback loop is closed around the input summing amplifier; the finite GBW product of this op amp causes a slight Q enhancement. If this is a problem, connect a small capacitor (10 pF – 100 pF) across R4 to provide some phase lead.

FIGURE 10. MODE 3

2.0 Modes of Operation (Continued)

MODE 3a: HP, BP, LP and Notch with External Op Amp (See *Figure 11*)

$$f_O = \frac{f_{CLK}}{100} \times \sqrt{\frac{R2}{R4}} \text{ or } \frac{f_{CLK}}{50} \times \sqrt{\frac{R2}{R4}}$$

$$Q = \sqrt{\frac{R2}{R4}} \times \frac{R3}{R2}$$

$$H_{OHP} = -\frac{R2}{R1}$$

$$H_{OBP} = -\frac{R3}{R1}$$

$$H_{OLP} = -\frac{R4}{R1}$$

f_n = notch frequency $= \frac{f_{CLK}}{100}\sqrt{\frac{R_h}{R_l}}$ or $\frac{f_{CLK}}{50}\sqrt{\frac{R_h}{R_l}}$

H_{ON} = gain of notch at

$$f = f_O = \left| Q\left(\frac{R_g}{R_l} H_{OLP} - \frac{R_g}{R_h} H_{OHP}\right)\right|$$

H_{n1} = gain of notch (as $f \rightarrow 0$) $= \frac{R_g}{R_l} \times H_{OLP}$

H_{n2} = gain of notch $\left(\text{as } f \rightarrow \frac{f_{CLK}}{2}\right)$

$$= -\frac{R_g}{R_h} \times H_{OHP}$$

MODE 4: Allpass, Bandpass, Lowpass Outputs (See *Figure 12*)

f_O = center frequency

$$= \frac{f_{CLK}}{100} \text{ or } \frac{f_{CLK}}{50};$$

f_z* = center frequency of the complex zero $\approx f_O$

$$Q = \frac{f_O}{BW} = \frac{R3}{R2};$$

Q_Z = quality factor of complex zero pair $= \frac{R3}{R1}$

For AP output make R1 = R2

H_{OAP}* = Allpass gain $\left(\text{at } 0 < f < \frac{f_{CLK}}{2}\right) = -\frac{R2}{R1} = -1$

H_{OLP} = Lowpass gain (as $f \rightarrow 0$)

$$= -\left(\frac{R2}{R1} + 1\right) = -2$$

H_{OBP} = Bandpass gain (at $f = f_O$)

$$= -\frac{R3}{R2}\left(1 + \frac{R2}{R1}\right) = -2\left(\frac{R3}{R2}\right)$$

Circuit Dynamics: $H_{OBP} = (H_{OLP}) \times Q = (H_{OAP} + 1)Q$

*Due to the sampled data nature of the filter, a slight mismatch of f_z and f_O occurs causing a 0.4 dB peaking around f_O of the allpass filter amplitude response (which theoretically should be a straight line). If this is unacceptable, Mode 5 is recommended.

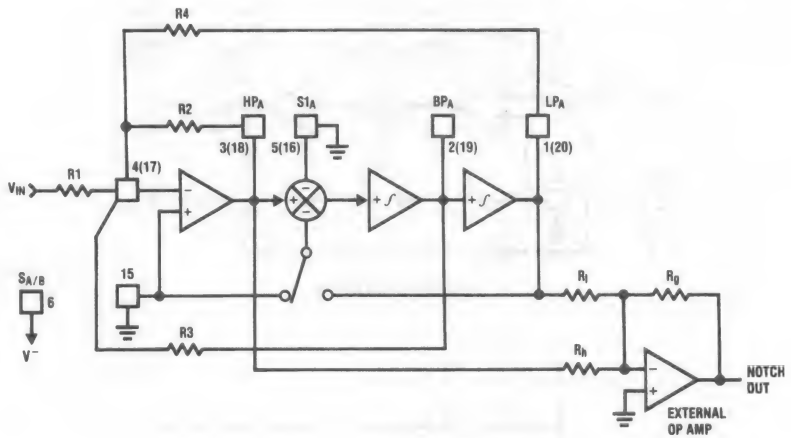

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FIGURE 11. MODE 3a

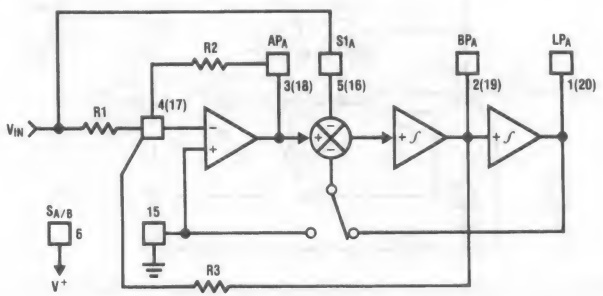

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FIGURE 12. MODE 4

2.0 Modes of Operation (Continued)

MODE 5: Numerator Complex Zeros, BP, LP (See *Figure 13*)

$$f_O = \sqrt{1+\frac{R2}{R4}} \times \frac{f_{CLK}}{100} \text{ or } \sqrt{1+\frac{R2}{R4}} \times \frac{f_{CLK}}{50}$$

$$f_z = \sqrt{1-\frac{R2}{R4}} \times \frac{f_{CLK}}{100} \text{ or } \sqrt{1-\frac{R1}{R4}} \times \frac{f_{CLK}}{50}$$

$$Q = \sqrt{1+R2/R4} \times \frac{R3}{R2}$$

$$Q_Z = \sqrt{1-R1/R4} \times \frac{R3}{R1}$$

H_{0z1} = gain at C.Z. output (as $f \rightarrow 0$ Hz) $= \dfrac{-R2(R4-R1)}{R1(R2+R4)}$

H_{0z2} = gain at C.Z. output $\left(\text{as } f \rightarrow \dfrac{f_{CLK}}{2}\right) = \dfrac{-R2}{R1}$

$$H_{OBP} = -\left(\frac{R2}{R1}+1\right) \times \frac{R3}{R2}$$

$$H_{OLP} = -\left(\frac{R2+R1}{R2+R4}\right) \times \frac{R4}{R1}$$

MODE 6a: Single Pole, HP, LP Filter (See *Figure 14*)

f_c = cutoff frequency of LP or HP output

$$= \frac{R2}{R3}\frac{f_{CLK}}{100} \text{ or } \frac{R2}{R3}\frac{f_{CLK}}{50}$$

$$H_{OLP} = -\frac{R3}{R1}$$

$$H_{OHP} = -\frac{R2}{R1}$$

MODE 6b: Single Pole LP Filter (Inverting and Non-Inverting) (See *Figure 15*)

f_c = cutoff frequency of LP outputs

$$\cong \frac{R2}{R3}\frac{f_{CLK}}{100} \text{ or } \frac{R2}{R3}\frac{f_{CLK}}{50}$$

H_{OLP1} = 1 (non-inverting)

$$H_{OLP2} = -\frac{R3}{R2}$$

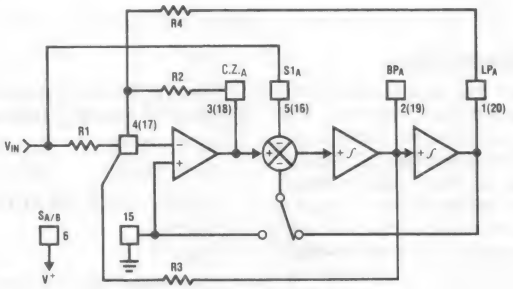

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FIGURE 13. MODE 5

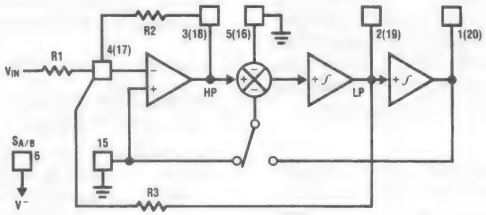

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FIGURE 14. MODE 6a

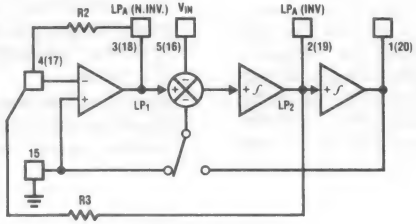

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FIGURE 15. MODE 6b

2.0 Modes of Operation (Continued)

TABLE I. Summary of Modes. Realizable filter types (e.g. low-pass) denoted by asterisks. Unless otherwise noted, gains of various filter outputs are inverting and adjustable by resistor ratios.

Mode	BP	LP	HP	N	AP	Number of Resistors	Adjustable f_{CLK}/f_O	Notes
1	*	*		*		3	No	
1a	(2) $H_{OBP1} = -Q$ $H_{OBP2} = +1$	$H_{OLP} + 1$				2	No	May need input buffer. Poor dynamics for high Q.
2	*	*		*		3	Yes (above $f_{CLK}/50$ or $f_{CLK}/100$)	
3	*	*	*			4	Yes	Universal State-Variable Filter. Best general-purpose mode.
3a	*	*	*	*		7	Yes	As above, but also includes resistor-tuneable notch.
4	*	*			*	3	No	Gives Allpass response with $H_{OAP} = -1$ and $H_{OLP} = -2$.
5	*	*			*	4		Gives flatter allpass response than above if $R_1 = R_2 = 0.02R_4$.
6a		*	*			3		Single pole.
6b		(2) $H_{OLP1} = +1$ $H_{OLP2} = \frac{-R3}{R2}$				2		Single Pole.

3.0 Applications Information

The MF10 is a general-purpose dual second-order state variable filter whose center frequency is proportional to the frequency of the square wave applied to the clock input (f_{CLK}). By connecting pin 12 to the appropriate DC voltage, the filter center frequency f_O can be made equal to either $f_{CLK}/100$ or $f_{CLK}/50$. f_O can be very accurately set (within ±6%) by using a crystal clock oscillator, or can be easily varied over a wide frequency range by adjusting the clock frequency. If desired, the f_{CLK}/f_O ratio can be altered by external resistors as in *Figures 9, 10, 11, 13, 14* and *15*. The filter Q and gain are determined by external resistors.

All of the five second-order filter types can be built using either section of the MF10. These are illustrated in *Figures 1* through *5* along with their transfer functions and some related equations. *Figure 6* shows the effect of Q on the shapes of these curves. When filter orders greater than two are desired, two or more MF10 sections can be cascaded.

3.1 DESIGN EXAMPLE

In order to design a second-order filter section using the MF10, we must define the necessary values of three parameters: f_0, the filter section's center frequency; H_0, the passband gain; and the filter's Q. These are determined by the characteristics required of the filter being designed.

As an example, let's assume that a system requires a fourth-order Chebyshev low-pass filter with 1 dB ripple, unity gain at DC, and 1000 Hz cutoff frequency. As the system order is four, it is realizable using both second-order sections of an MF10. Many filter design texts include tables that list the characteristics (f_O and Q) of each of the second-order filter sections needed to synthesize a given higher-order filter. For the Chebyshev filter defined above, such a table yields the following characteristics:

$f_{0A} = 529$ Hz $\quad Q_A = 0.785$

$f_{0B} = 993$ Hz $\quad Q_B = 3.559$

For unity gain at DC, we also specify:

$H_{0A} = 1$

$H_{0B} = 1$

The desired clock-to-cutoff-frequency ratio for the overall filter of this example is 100 and a 100 kHz clock signal is available. Note that the required center frequencies for the two second-order sections will not be obtainable with clock-to-center-frequency ratios of 50 or 100. It will be necessary to adjust $\frac{f_{CLK}}{f_0}$ externally. From Table I, we see that Mode 3 can be used to produce a low-pass filter with resistor-adjustable center frequency.

In most filter designs involving multiple second-order stages, it is best to place the stages with lower Q values ahead of stages with higher Q, especially when the higher Q is greater than 0.707. This is due to the higher relative gain at the center frequency of a higher-Q stage. Placing a stage with lower Q ahead of a higher-Q stage will provide some attenuation at the center frequency and thus help avoid clipping of signals near this frequency. For this example, stage A has the lower Q (0.785) so it will be placed ahead of the other stage.

For the first section, we begin the design by choosing a convenient value for the input resistance: $R_{1A} = 20$k. The absolute value of the passband gain H_{OLPA} is made equal

3.0 Applications Information (Continued)

to 1 by choosing R_{4A} such that: $R_{4A} = -H_{OLPA}\, R_{1A} = R_{1A} = 20k$. If the 50/100/CL pin is connected to mid-supply for nominal 100:1 clock-to-center-frequency ratio, we find R_{2A} by:

$$R_{2A} = R_{4A}\frac{f_{0A}^2}{(f_{CLK}/100)^2} = 2 \times 10^4 \times \frac{(529)^2}{(1000)^2} = 5.6k \text{ and}$$

$$R_{3A} = Q_A\sqrt{R_{2A}R_{4A}} = 0.785\sqrt{5.6 \times 10^3 \times 2 \times 10^4} = 8.3k$$

The resistors for the second section are found in a similar fashion:

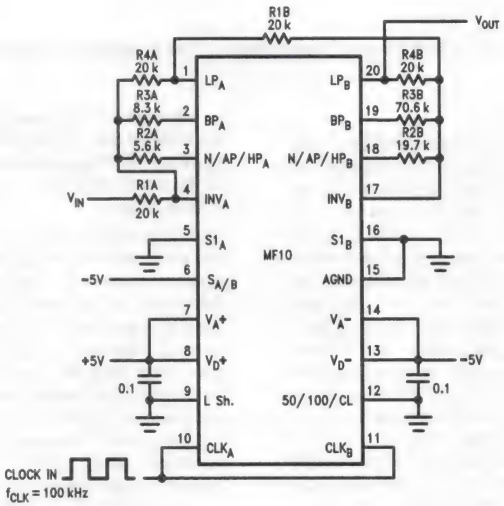
$R_{1B} = 20k$

$R_{4B} = R_{1B} = 20k$

$$R_{2B} = R_{4B}\frac{f_{0B}^2}{(f_{CLK}/100)^2} = 20k\frac{(993)^2}{(1000)^2} = 19.7k$$

$$R_{3B} = Q_B\sqrt{R_{2B}R_{4B}} = 3.559\sqrt{1.97\times10^4\times2\times10^4} = 70.6k$$

The complete circuit is shown in *Figure 16* for split ±5V power supplies. Supply bypass capacitors are highly recommended.



TL/H/10399–25

FIGURE 16. Fourth-Order Chebyshev Low-Pass Filter from Example in 3.1. ±5V Power Supply. 0V–5V TTL or −5V ±5V CMOS Logic Levels.

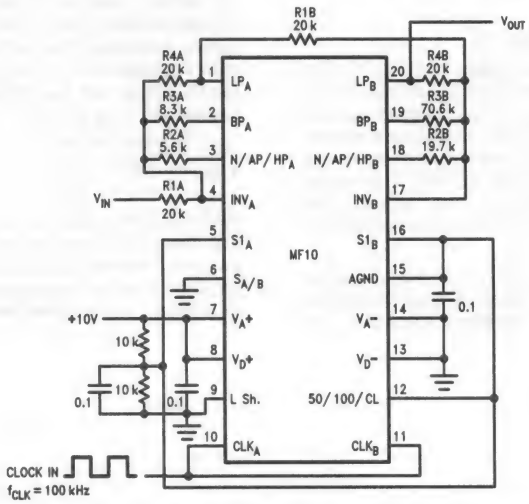


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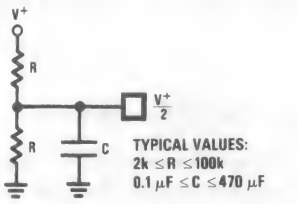
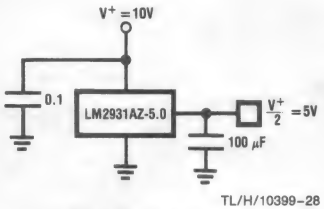
FIGURE 17. Fourth-Order Chebyshev Low-Pass Filter from Example in 3.1. Single +10V Power Supply. 0V–5V TTL Logic Levels. Input Signals Should be Referred to Half-Supply or Applied through a Coupling Capacitor.

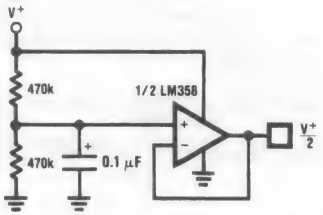
3.0 Applications Information (Continued)



(a) Resistive Divider with Decoupling Capacitor



(b) Voltage Regulator



(c) Operational Amplifier with Divider

FIGURE 18. Three Ways of Generating $\frac{V^+}{2}$ for Single-Supply Operation

3.2 SINGLE SUPPLY OPERATION

The MF10 can also operate with a single-ended power supply. *Figure 17* shows the example filter with a single-ended power supply. V_A^+ and V_D^+ are again connected to the positive power supply (8V to 14V), and V_A^- and V_D^- are connected to ground. The A_{GND} pin must be tied to $V^+/2$ for single supply operation. This half-supply point should be very "clean", as any noise appearing on it will be treated as an input to the filter. It can be derived from the supply voltage with a pair of resistors and a bypass capacitor *(Figure 18a)*, or a low-impedance half-supply voltage can be made using a three-terminal voltage regulator or an operational amplifier *(Figures 18b* and *18c)*. The passive resistor divider with a bypass capacitor is sufficient for many applications, provided that the time constant is long enough to reject any power supply noise. It is also important that the half-supply reference present a low impedance to the clock frequency, so at very low clock frequencies the regulator or op-amp approaches may be preferable because they will require smaller capacitors to filter the clock frequency. The main power supply voltage should be clean (preferably regulated) and bypassed with 0.1 μF.

3.3 DYNAMIC CONSIDERATIONS

The maximum signal handling capability of the MF10, like that of any active filter, is limited by the power supply voltages used. The amplifiers in the MF10 are able to swing to within about 1V of the supplies, so the input signals must be kept small enough that none of the outputs will exceed these limits. If the MF10 is operating on ±5V, for example, the outputs will clip at about 8 V_{p-p}. The maximum input voltage multiplied by the filter gain should therefore be less than 8 V_{p-p}.

Note that if the filter Q is high, the gain at the lowpass or highpass outputs will be much greater than the nominal filter gain *(Figure 6)*. As an example, a lowpass filter with a Q of 10 will have a 20 dB peak in its amplitude response at f_O. If the nominal gain of the filter H_{OLP} is equal to 1, the gain at f_O will be 10. The maximum input signal at f_O must therefore be less than 800 mV_{p-p} when the circuit is operated on ±5V supplies.

Also note that one output can have a reasonable small voltage on it while another is saturated. This is most likely for a circuit such as the notch in Mode 1 *(Figure 7)*. The notch output will be very small at f_O, so it might appear safe to apply a large signal to the input. However, the bandpass will have its maximum gain at f_O and can clip if overdriven. If one output clips, the performance at the other outputs will be degraded, so avoid overdriving any filter section, even ones whose outputs are not being directly used. Accompanying *Figures 7* through *15* are equations labeled "circuit dynamics", which relate the Q and the gains at the various outputs. These should be consulted to determine peak circuit gains and maximum allowable signals for a given application.

3.4 OFFSET VOLTAGE

The MF10's switched capacitor integrators have a higher equivalent input offset voltage than would be found in a typical continuous-time active filter integrator. *Figure 19* shows an equivalent circuit of the MF10 from which the output DC offsets can be calculated. Typical values for these offsets with $S_{A/B}$ tied to V^+ are:

V_{os1} = opamp offset = ±5 mV

V_{os2} = −150 mV @ 50:1 −300 mV @ 100:1

V_{os3} = −70 mV @ 50:1 −140 mV @ 100:1

When $S_{A/B}$ is tied to V^-, V_{os2} will approximately halve. The DC offset at the BP output is equal to the input offset of the lowpass integrator (V_{os3}). The offsets at the other outputs depend on the mode of operation and the resistor ratios, as described in the following expressions.

3.0 Applications Information (Continued)

Mode 1 and Mode 4

$$V_{OS(N)} = V_{OS1}\left(\frac{1}{Q} + 1 \left\| H_{OLP} \right\|\right) - \frac{V_{OS3}}{Q}$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N)} - V_{OS2}$$

Mode 1a

$$V_{OS(N.INV.BP)} = \left(1 + \frac{1}{Q}\right) V_{OS1} - \frac{V_{OS3}}{Q}$$

$$V_{OS(INV.BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N.INV.BP)} - V_{OS2}$$

Mode 2 and Mode 5

$$V_{OS(N)} = \left(\frac{R2}{R_p} + 1\right) V_{OS1} \times \frac{1}{1 + R2/R4} + V_{OS2}\frac{1}{1 + R4/R2} - \frac{V_{OS3}}{Q\sqrt{1 + R2/R4}}$$

$$R_p = R1//R3//R4$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS(N)} - V_{OS2}$$

Mode 3

$$V_{OS(HP)} = V_{OS2}$$

$$V_{OS(BP)} = V_{OS3}$$

$$V_{OS(LP)} = V_{OS1}\left[1 + \frac{R4}{R_p}\right] - V_{OS2}\left(\frac{R4}{R2}\right) - V_{OS3}\left(\frac{R4}{R3}\right)$$

$$Rp = R1//R2//R3$$

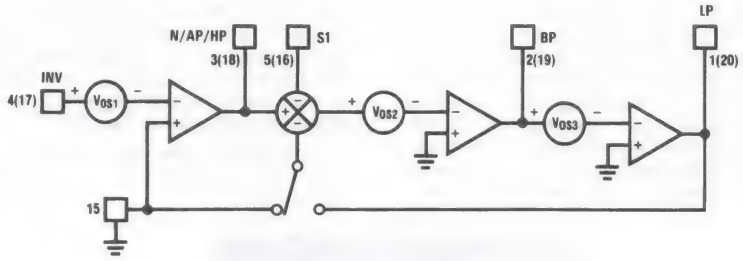


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FIGURE 19. MF10 Offset Voltage Sources

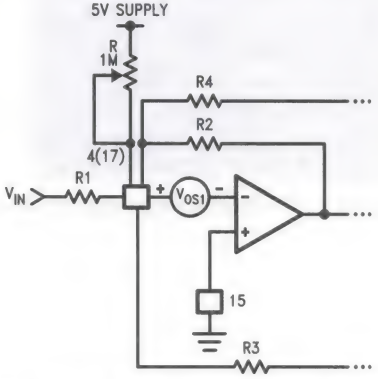

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FIGURE 20. Method for Trimming V_{OS}

3.0 Applications Information (Continued)

For most applications, the outputs are AC coupled and DC offsets are not bothersome unless large signals are applied to the filter input. However, larger offset voltages will cause clipping to occur at lower AC signal levels, and clipping at any of the outputs will cause gain nonlinearities and will change f_O and Q. When operating in Mode 3, offsets can become excessively large if R2 and R4 are used to make f_{CLK}/f_O significantly higher than the nominal value, especially if Q is also high. An extreme example is a bandpass filter having unity gain, a Q of 20, and $f_{CLK}/f_O = 250$ with pin 12 tied to ground (100:1 nominal). R4/R2 will therefore be equal to 6.25 and the offset voltage at the lowpass output will be about +1V. Where necessary , the offset voltage can be adjusted by using the circuit of *Figure 20*. This allows adjustment of V_{OS1}, which will have varying effects on the different outputs as described in the above equations. Some outputs cannot be adjusted this way in some modes, however ($V_{OS(BP)}$ in modes 1a and 3, for example).

3.5 SAMPLED DATA SYSTEM CONSIDERATIONS

The MF10 is a sampled data filter, and as such, differs in many ways from conventional continuous-time filters. An important characteristic of sampled-data systems is their effect on signals at frequencies greater than one-half the sampling frequency. (The MF10's sampling frequency is the same as its clock frequency.) If a signal with a frequency greater than one-half the sampling frequency is applied to the input of a sampled data system, it will be "reflected" to a frequency less than one-half the sampling frequency. Thus, an input signal whose frequency is $f_s/2 + 100$ Hz will cause the system to respond as though the input frequency was $f_s/2 - 100$ Hz. This phenomenon is known as "aliasing", and can be reduced or eliminated by limiting the input signal spectrum to less than $f_s/2$. This may in some cases require the use of a bandwidth-limiting filter ahead of the MF10 to limit the input spectrum. However, since the clock frequency is much higher than the center frequency, this will often not be necessary.

Another characteristic of sampled-data circuits is that the output signal changes amplitude once every sampling period, resulting in "steps" in the output voltage which occur at the clock rate *(Figure 21)*. If necessary, these can be "smoothed" with a simple R–C low-pass filter at the MF10 output.

The ratio of f_{CLK} to f_C (normally either 50:1 or 100:1) will also affect performance. A ratio of 100:1 will reduce any aliasing problems and is usually recommended for wideband input signals. In noise sensitive applications, however, a ratio of 50:1 may be better as it will result in 3 dB lower output noise. The 50:1 ratio also results in lower DC offset voltages, as discussed in Section 3.4.

The accuracy of the f_{CLK}/f_O ratio is dependent on the value of Q. This is illustrated in the curves under the heading "Typical Performance Characteristics". As Q is changed, the true value of the ratio changes as well. Unless the Q is low, the error in f_{CLK}/f_O will be small. If the error is too large for a specific application, use a mode that allows adjustment of the ratio with external resistors.

It should also be noted that the product of Q and f_O should be limited to 300 kHz when $f_O < 5$ kHz, and to 200 kHz for $f_O > 5$ kHz.

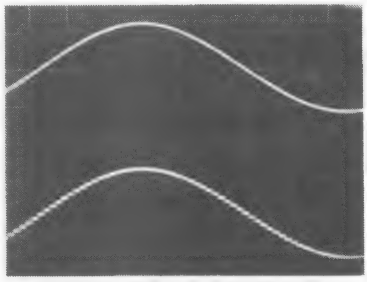

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FIGURE 21. The Sampled-Data Output Waveform

Section 2

Analog Switches/ Multiplexers

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Analog Switches/Multiplexers Selection Guide 2-4
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Analog Switch
Definition of Terms

R_{ON}: Resistance between the output and the input of an addressed channel.

I_S: Current at any switch input. This is leakage current when the switch is ON.

I_D: Current at any switch input going into the switch. This is leakage current when the switch is OFF.

C_S: Capacitance between any open terminal "S" and ground.

C_D: Capacitance between any open terminal "D" and ground.

$I_D - I_S$: Leakage current that flows from the closed switch into the body. This leakage is the difference between the current I_D going into the switch and the current I_S going out of the switch.

t_{RAN}: Delay time when switching from one address state to another.

t_{ON}: Delay time between the 50% points of an enable input and the switch ON condition.

t_{OFF}: Delay time between the 50% points of the enable input and the switch OFF condition.

National Semiconductor

Analog Switch/Multiplexer Selection Guide

Part Number	Function	Logic Input	V_S (Typ)	T_{ON}/T_{OFF} ns (Typ)	R_{ON} Ω
AH0014	DPDT	TTL, DTL	+10/−22	350/600	75
AH0015	QUAD SPST	TTL, DTL	+10/−22	100/600	75
AH0019	DUAL DPST	TTL, DTL	+10/−22	100/600	75
AH5011	QUAD SPST	TTL, CMOS	—	150/300	100
AH5012		TTL, CMOS	—	150/300	150
CD4016		CMOS	±7.5	20/40	850
CD4066		CMOS	±7.5	25/50	280
LF11201/LF13201		TTL	±15	90/500	200
LF11202/LF13202		TTL	±15	90/500	200
LF11331/LF13331		TTL	±15	90/500	200
LF11332/LF13332		TTL	±15	90/500	200
LF11333/LF13333		TTL	±15	90/500	200
MM74HC4016		CMOS	±12	5/8	40
AH5020	DUAL SPDT	TTL, CMOS	—	150/300	150
CD4053	TRIPLE SPDT	CMOS	±7.5	160/75	300
MM74HC4053		CMOS	±6.0	15/16	40
AH5009	4-CHANNEL	TTL, CMOS	—	150/300	100
AH5010		TTL, CMOS	—	150/300	150
CD4052	4-CHANNEL	CMOS	±7.5	160/75	300
CD4529B	DIFFERENTIAL	CMOS	±7.5	50	350
LF13509		TTL, CMOS	±18	1600/200	350
MM74HC4052		CMOS	±6.0	15/16	40
CD4051	8-CHANNEL	CMOS	±7.5	160/75	300
CD4529B		CMOS	±7.5	50	350
LF13508		TTL, CMOS	±18	1600/200	350
MM74HC4051		CMOS	±6.0	15/16	40

National Semiconductor

AH0014/AH0014C* DPDT, AH0015/AH0015C Quad SPST, AH0019/AH0019C* Dual DPST-TTL/DTL Compatible MOS Analog Switches

General Description

This series of TTL/DTL compatible MOS analog switches feature high speed with internal level shifting and driving. The package contains two monolithic integrated circuit chips: the MOS analog chip is similar to the MM450 type which consists of four MOS analog switch transistors; the second chip is a bipolar I.C. gate and level shifter. The series is available in hermetic dual-in-line package.

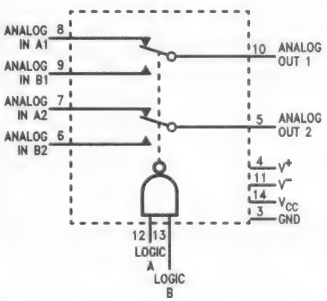
These switches are particularly suited for use in both military and industrial applications such as commutators in data acquisition systems, multiplexers, A/D and D/A converters, long time constant integrators, sample and hold circuits, modulators/demodulators, and other analog signal switching applications.

The AH0014, AH0015 and AH0019 are specified for operation over the −55°C to +125°C military temperature range. The AH0014C, AH0015C and AH0019C are specified for operation over the −25°C to +85°C temperature range.

Features

- Large analog voltage switching ±10V
- Fast switching speed 500 ns
- Operation over wide range of power supplies
- Low ON resistance 200Ω
- High OFF resistance $10^{11}\Omega$
- Analog signals in excess of 25 MHz
- Fully compatible with DTL or TTL logic
- Includes gating and level shifting

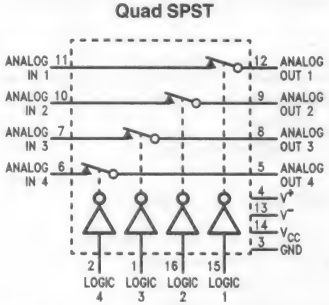
Block and Connection Diagrams



Note: All logic inputs shown at logic "1". TL/K/10125–1

Order Number AH0014D or AH0014CD
See NS Package Number D14D

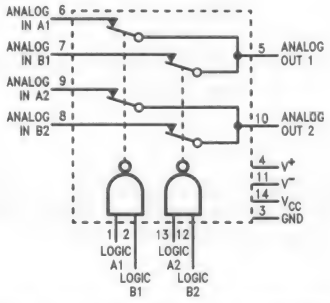
Quad SPST



Note: All logic inputs shown at logic "1". TL/K/10125–2

Order Number AH0015D or AH0015CD
See NS Package Number D16C

Dual DPST



Note: All logic inputs shown at logic "1".

TL/K/10125–3

Order Number AH0019D or AH0019CD
See NS Package Number D14D

*Previously called NH0014/NH0014C and NH0019/NH0019C

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

V_{CC} Supply Voltage	7.0V
V^- Supply Voltage	−30V
V^+ Supply Voltage	+30V
V^+/V^- Voltage Differential	40V
Logic Input Voltage	5.5V
Storage Temperature Range	−65°C to +150°C
Operating Temperature Range	
AH0014, AH0015, AH0019	−55°C to +125°C
AH0014C, AH0015C, AH0019C	−25°C to +85°C
Lead Temperature (Soldering, 10 sec)	300°C

Electrical Characteristics (Notes 1 and 2)

Parameter	Conditions	Min	Typ	Max	Units
Logical "1" Input Voltage	V_{CC} = 4.5V	2.0			V
Logical "0" Input Voltage	V_{CC} = 4.5V			0.8	V
Logical "1" Input Current	V_{CC} = 5.5V, V_{IN} = 2.4V			5	μA
Logical "1" Input Current	V_{CC} = 5.5V, V_{IN} = 5.5V			1	μA
Logical "0" Input Current	V_{CC} = 5.5V, V_{IN} = 0.4V		0.2	0.4	mA
Power Supply Current Logical "1" Input—Each Gate (Note 3)	V_{CC} = 5.5V, V_{IN} = 4.5V		0.85	1.6	mA
Power Supply Current Logical "0" Input—Each Gate (Note 3)	V_{CC} = 5.5V, V_{IN} = 0V				
AH0014, AH0014C			1.5	3.0	mA
AH0015, AH0015C			0.22	0.41	mA
AH0019, AH0019C			0.22	0.41	mA
Analog Switch ON Resistance—Each Gate	V_{IN} (Analog) = +10V		75	200	Ω
	V_{IN} (Analog) = −10V		150	600	Ω
Analog Switch OFF Resistance			10^{11}		Ω
Analog Switch Input Leakage Current—Each Input (Note 4)	V_{IN} = −10V				
AH0014, AH0015, AH0019	T_A = 25°C		25	200	pA
	T_A = 125°C		25	200	nA
AH0014C, AH0015C, AH0019C	T_A = 25°C		0.1	10	nA
	T_A = 70°C		30	100	nA
Analog Switch Output Leakage Current—Each Output (Note 4)	V_{OUT} = −10V				
AH0014, AH0015, AH0019	T_A = 25°C		40	400	pA
	T_A = 125°C		40	400	nA
AH0014C, AH0015C, AH0019C	T_A = 25°C		0.05	10	nA
	T_A = 70°C		4	50	nA
Analog Input (Drain) Capacitance	1 MHz @ Zero Bias		8	10	pF
Output Source Capacitance	1 MHz @ Zero Bias		11	13	pF
Analog Turn-OFF Time—t_{OFF}	See Test Circuit; T_A = 25°C		600	750	ns
Analog Turn-ON Time—t_{ON}	See Test Circuit; T_A = 25°C				
AH0014, AH0014C			350	425	ns
AH0015, AH0015C			100	150	ns
AH0019, AH0019C			100	150	ns

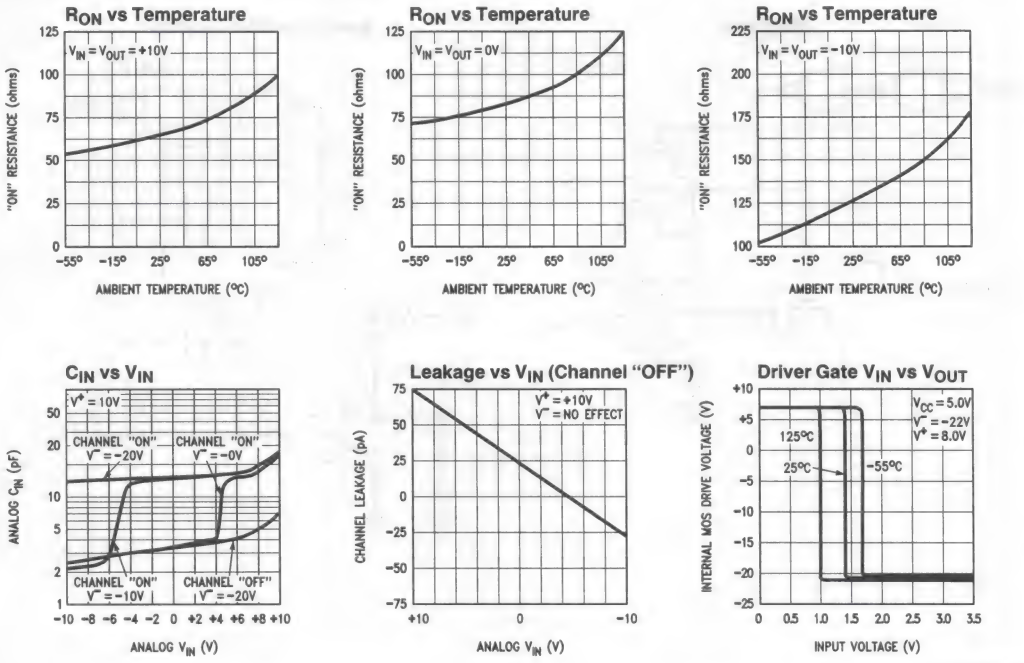
Note 1: Min/max limits apply across the guaranteed temperature range of −55°C to +125°C for AH0014, AH0015, AH0019 and −25°C to +85°C for AH0014C, AH0015C, AH0019C. V^- = −20V. V^+ = +10V and an analog test current of 1 mA unless otherwise specified.

Note 2: All typical values are measured at T_A = 25°C with V_{CC} = 5.0V. V^+ = +10V, V^- = −22V.

Note 3: Current measured is drawn from V_{CC} supply.

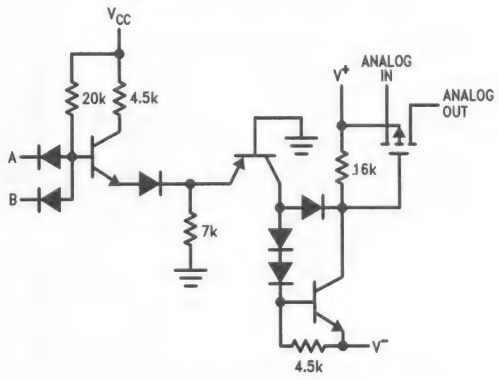
Note 4: All analog switch pins except measurement pin are tied to V^+.

Analog Switch Characteristics (Note 2)



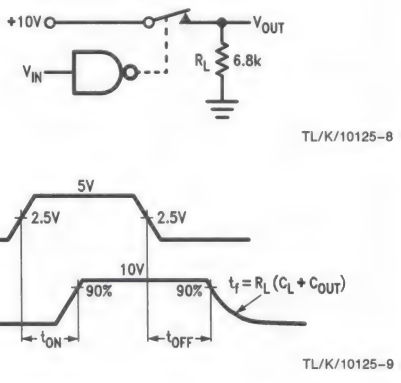
TL/K/10125–6

Schematic (Single Driver Gate and MOS Switch Shown)



TL/K/10125–7

Analog Switching Time Test Circuit

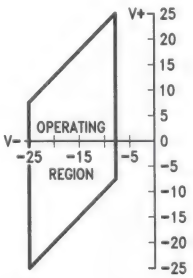


TL/K/10125–8

TL/K/10125–9

Selecting Power Supply Voltage

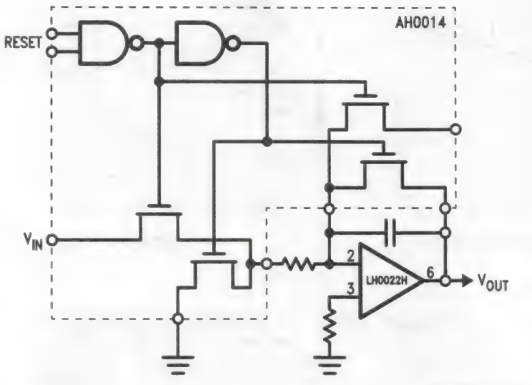
The graph shows the boundary conditions which must be used for proper operation of the unit. The range of operation for power supply V^- is shown on the X axis. It must be between $-25V$ and $-8V$. The allowable range for power supply V^+ is governed by supply V^-. With a value chosen for V^-, V^+ may be selected as any value along a vertical line passing through the V^- value and terminated by the boundaries of the operating region. A voltage difference between power supplies of at least 5V should be maintained for adequate signal swing.



TL/K/10125–10

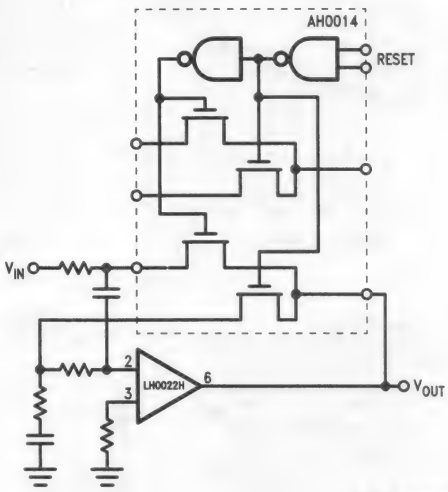
Typical Applications

Integrator



TL/K/10125–4

Reset Stabilized Amplifier



TL/K/10125–5

National Semiconductor

AH5009/AH5010/AH5011/AH5012 Monolithic Analog Current Switches

General Description

A versatile family of monolithic JFET analog switches economically fulfills a wide variety of multiplexing and analog switching applications.

Even numbered switches may be driven directly from standard 5V logic, whereas the odd numbered switches are intended for applications utilizing 10V or 15V logic. The monolithic construction guarantees tight resistance match and track.

For voltage switching applications see LF13331, LF13332, and LF13333 Analog Switch Family, or the CMOS Analog Switch Family.

Applications

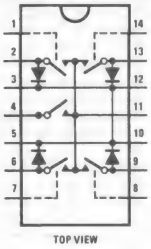
- A/D and D/A converters
- Micropower converters
- Industrial controllers
- Position controllers
- Data acquisition
- Active filters
- Signal multiplexers/demultiplexers
- Multiple channel AGC
- Quad compressors/expanders
- Choppers/demodulators
- Programmable gain amplifiers
- High impedance voltage buffer
- Sample and hold

Features

■ Interfaces with standard TTL and CMOS	
■ "ON" resistance match	2Ω
■ Low "ON" resistance	100Ω
■ Very low leakage	50 pA
■ Large analog signal range	±10V peak
■ High switching speed	150 ns
■ Excellent isolation between channels	80 dB at 1 kHz

Connection and Schematic Diagrams (All switches shown are for logical "1" input)

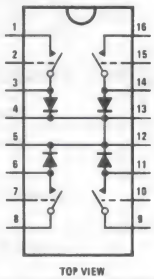
Dual-In-Line Package



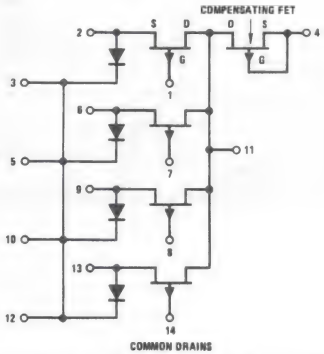
LOGIC DRIVE	4 CHANNEL MUX	4 SPST SWITCHES
5V LOGIC	AH5010C	AH5012C
15V LOGIC	AH5009C	AH5011C

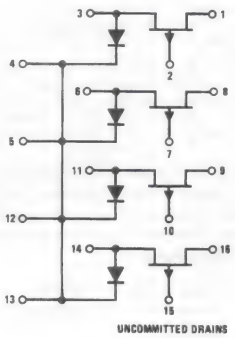
AH5009C and AH5010C MUX Switches
(4-Channel Version Shown)
Order Number AH5009CM,
AH5009CN, AH5010CM or AH5010CN
See NS Package Number M14A or N14A

Dual-In-Line Package



AH5011C and AH5012C SPST Switches
(Quad Version Shown)
Order Number AH5011CM,
AH5011CN, AH5012CM or AH5012CN
See NS Package Number M16A or N16A





Note: All diode cathodes are internally connected to the substrate.

TL/H/5659–1

2

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Input Voltage	
AH5009/AH5010/AH5011/AH5012	30V
Positive Analog Signal Voltage	30V
Negative Analog Signal Voltage	−15V
Diode Current	10 mA
Drain Current	30 mA
Soldering Information:	
N Package 10 sec	300°C
SO Package Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C
Power Dissipation	500 mW
Operating Temperature Range	−25°C to +85°C
Storage Temperature Range	−65°C to +150°C

Electrical Characteristics AH5010 and AH5012 (Notes 2 and 3)

Symbol	Parameter	Conditions	Typ	Max	Units
I_{GSX}	Input Current "OFF"	$4.5V \le V_{GD} \le 11V$, $V_{SD}=0.7V$ $T_A=85°C$	0.01	0.2 10	nA nA
$I_{D(OFF)}$	Leakage Current "OFF"	$V_{SD}=0.7V$, $V_{GS}=3.8V$ $T_A=85°C$	0.02	0.2 10	nA nA
$I_{G(ON)}$	Leakage Current "ON"	$V_{GD}=0V$, $I_S=1$ mA $T_A=85°C$	0.08	1 200	nA nA
$I_{G(ON)}$	Leakage Current "ON"	$V_{GD}=0V$, $I_S=2$ mA $T_A=85°C$	0.13	5 10	nA μA
$I_{G(ON)}$	Leakage Current "ON"	$V_{GD}=0V$, $I_S=-2$ mA $T_A=85°C$	0.1	10 20	nA μA
$r_{DS(ON)}$	Drain-Source Resistance	$V_{GS}=0.35V$, $I_S=2$ mA $T_A=+85°C$	90	150 240	Ω Ω
V_{DIODE}	Forward Diode Drop	$I_D=0.5$ mA		0.8	V
$r_{DS(ON)}$	Match	$V_{GS}=0V$, $I_D=1$ mA	4	20	Ω
T_{ON}	Turn "ON" Time	See AC Test Circuit	150	500	ns
T_{OFF}	Turn "OFF" Time	See AC Test Circuit	300	500	ns
CT	Cross Talk	See AC Test Circuit	120		dB

Electrical Characteristics AH5009 and AH5011 (Notes 2 and 3)

Symbol	Parameter	Conditions	Typ	Max	Units
I_{GSX}	Input Current "OFF"	$11V \le V_{GD} \le 15V$, $V_{SD}=0.7V$ $T_A=85°C$	0.01	0.2 10	nA nA
$I_{D(OFF)}$	Leakage Current "OFF"	$V_{SD}=0.7V$, $V_{GS}=10.3V$ $T_A=85°C$	0.01	0.2 10	nA nA
$I_{G(ON)}$	Leakage Current "ON"	$V_{GD}=0V$, $I_S=1$ mA $T_A=85°C$	0.04	0.5 100	nA nA
$I_{G(ON)}$	Leakage Current "ON"	$V_{GD}=0V$, $I_S=2$ mA $T_A=85°C$		2 1	nA μA
$I_{G(ON)}$	Leakage Current "ON"	$V_{GD}=0V$, $I_S=-2$ mA $T_A=85°C$		5 2	nA μA
$r_{DS(ON)}$	Drain-Source Resistance	$V_{GS}=1.5V$, $I_S=2$ mA $T_A=85°C$	60	100 160	Ω Ω
V_{DIODE}	Forward Diode Drop	$I_D=0.5$ mA		0.8	V
$r_{DS(ON)}$	Match	$V_{GS}=0V$, $I_D=1$ mA	2	10	Ω
T_{ON}	Turn "ON" Time	See AC Test Circuit	150	50	ns
T_{OFF}	Turn "OFF" Time	See AC Test Circuit	300	500	ns
CT	Cross Talk	See AC Test Circuit. f = 100 Hz.	120		dB

Note 1: Absolute maximum ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: Test conditions 25°C unless otherwise noted.

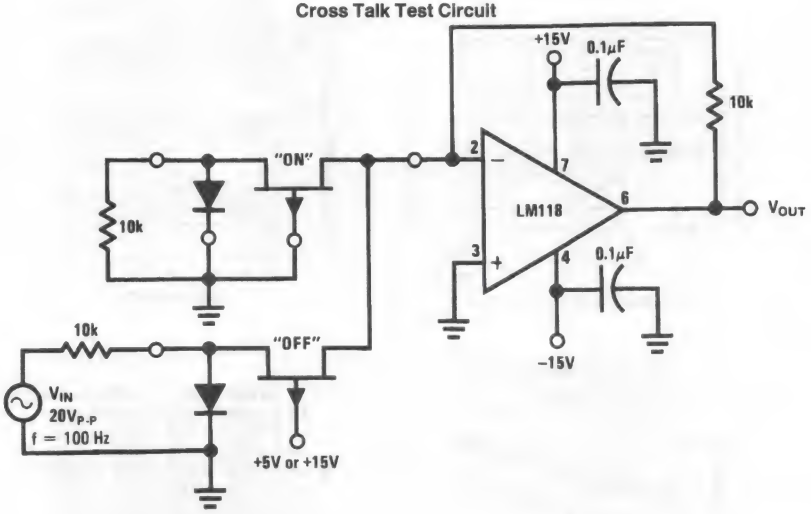
Note 3: "OFF" and "ON" notation refers to the conduction state of the FET switch.

Note 4: Thermal Resistance:

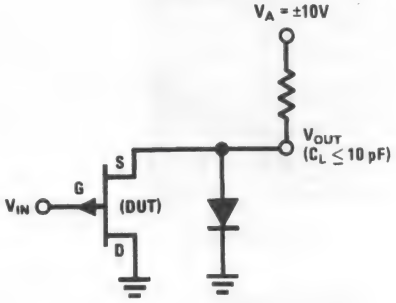
	θ_{JA}
N14A, N16A	92°C/W
M14A, M16A	115°C/W

Test Circuits and Switching Time Waveforms

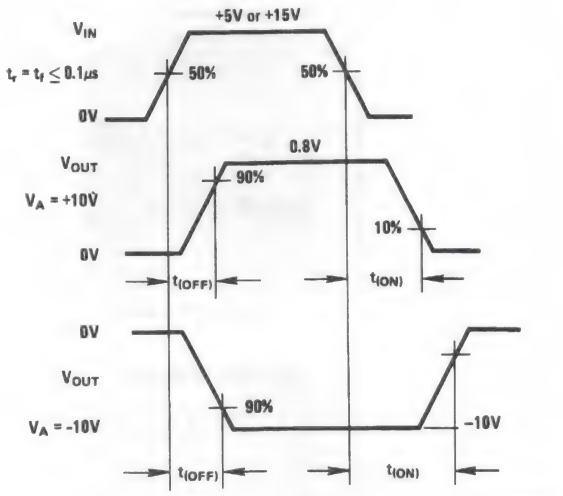
Cross Talk Test Circuit



AC Test Circuit



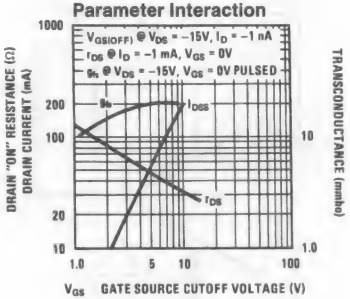
Time Waveforms



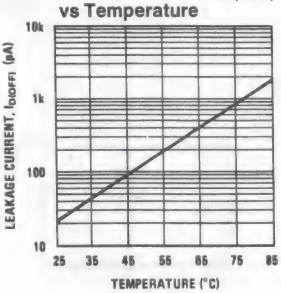
TL/H/5659–2

Typical Performance Characteristics

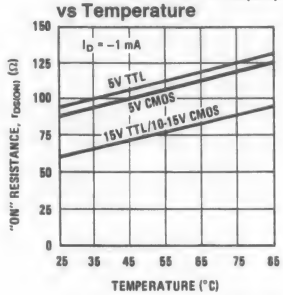
Parameter Interaction



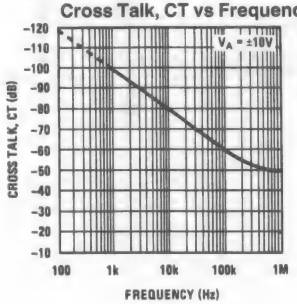
Leakage Current, $I_{D(OFF)}$ vs Temperature



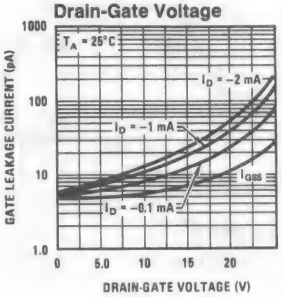
"ON" Resistance, $r_{DS(ON)}$ vs Temperature



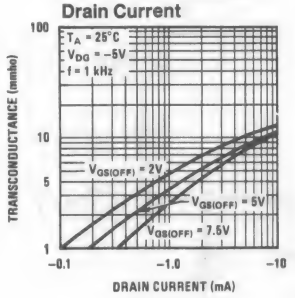
Cross Talk, CT vs Frequency



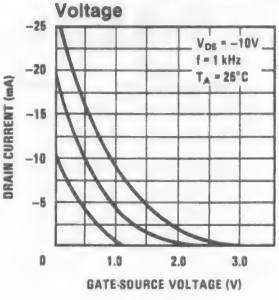
Leakage Current vs Drain-Gate Voltage



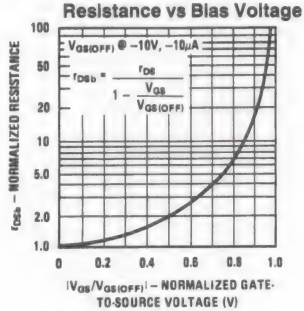
Transconductance vs Drain Current



Drain Current vs Bias Voltage



Normalized Drain Resistance vs Bias Voltage



TL/H/5659–3

Applications Information

Theory of Operation

The AH series of analog switches are primarily intended for operation in current mode switch applications; i.e., the drains of the FET switch are held at or near ground by operating into the summing junction of an operational amplifier. Limiting the drain voltage to under a few hundred millivolts eliminates the need for a special gate driver, allowing the switches to be driven directly by standard TTL (AH5010), 5V-10V CMOS (AH5010), open collector 15V TTL/CMOS (AH5009).

Two basic switch configurations are available: 4 independent switches (SPST) and 4 pole switches used for multiplexing (4 PST-MUX). The MUX versions such as the AH5009 offer common drains and include a series FET operated at V_{GS} = 0V. The additional FET is placed in the feedback path in order to compensate for the "ON" resistance of the switch FET as shown in *Figure 1*.

The closed-loop gain of *Figure 1* is:

$$A_{VCL} = \frac{R2 + r_{DS(ON)Q2}}{R1 + r_{DS(ON)Q1}}$$

For R1 = R2, gain accuracy is determined by the $r_{DS(ON)}$ match between Q1 and Q2. Typical match between Q1 and Q2 is 4 ohms resulting in a gain accuracy of 0.05% (for R1 = R2 = 10 kΩ).

Noise Immunity

The switches with the source diodes grounded exhibit improved noise immunity for positive analog signals in the "OFF" state. With V_{IN} = 15V and the V_A = 10V, the source of Q1 is clamped to about 0.7V by the diode (V_{GS} = 14.3V) ensuring that ac signals imposed on the 10V input will not gate the FET "ON."

Selection of Gain Setting Resistors

Since the AH series of analog switches are operated in current mode, it is generally advisable to make the signal current as large as possible. However, current through the FET switch tends to forward bias the source to gate junction and the signal shunting diode resulting in leakage through these junctions. As shown in *Figure 2*, $I_{G(ON)}$ represents a finite error in the current reaching the summing junction of the op amp.

Secondly, the $r_{DS(ON)}$ of the FET begins to "round" as I_S approaches I_{DSS}. A practical rule of thumb is to maintain I_S at less than $1/10$ of I_{DSS}.

Combining the criteria from the above discussion yields:

$$R1_{min} \geq \frac{V_{A(MAX)}\, A_D}{I_{G(ON)}} \quad (2a)$$

or:

$$\geq \frac{V_{A(MAX)}}{I_{DSS}/10} \quad (2b)$$

whichever is larger.

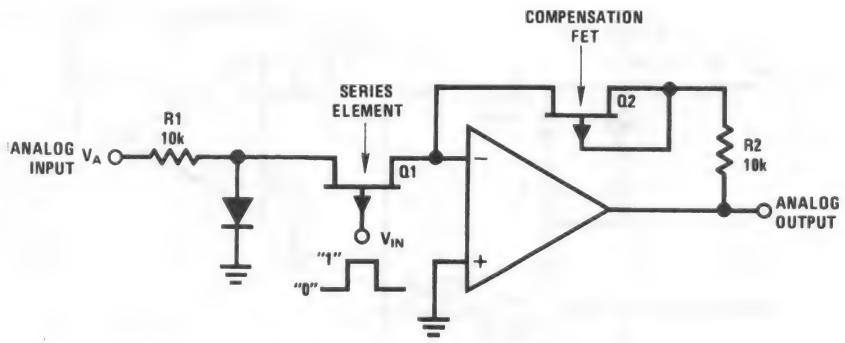
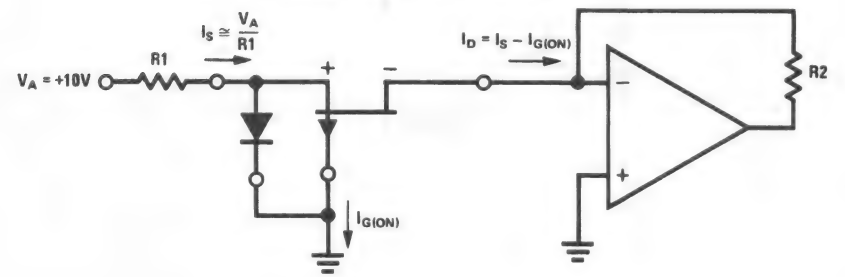


FIGURE 1. Use of Compensation FET



TL/H/5659–4

FIGURE 2. On Leakage Current, $I_{G(ON)}$

Applications Information (Continued)

Where: $V_{A(MAX)}$ = Peak amplitude of the analog input signal

A_D = Desired accuracy

$I_{G(ON)}$ = Leakage at a given I_S

I_{DSS} = Saturation current of the FET switch

$\cong 20$ mA

In a typical application, V_A might $= \pm 10V$, $A_D = 0.1\%$, $0°C \leq T_A \leq 85°C$. The criterion of equation (2b) predicts:

$$R1_{(MIN)} \geq \frac{(10V)}{\left(\frac{20\ mA}{10}\right)} = 5\ k\Omega$$

For R1 = 5k, $I_S \cong 10V/5k$ or 2 mA. The electrical characteristics guarantee an $I_{G(ON)} \leq 1\mu A$ at 85°C for the AH5010. Per the criterion of equation (2a):

$$R1_{(MIN)} \geq \frac{(10V)(10^{-3})}{1 \times 10^{-6}} \geq 10\ k\Omega$$

Since equation (2a) predicts a higher value, the 10k resistor should be used.

The "OFF" condition of the FET also affects gain accuracy. As shown in *Figure 3*, the leakage across Q2, $I_{D(OFF)}$ represents a finite error in the current arriving at the summing junction of the op amp.

Accordingly:

$$R1_{(MAX)} \leq \frac{V_{A(MIN)}\ A_D}{(N)\ I_{D(OFF)}}$$

Where: $V_{A(MIN)}$ = Minimum value of the analog input signal

A_D = Desired accuracy

N = Number of channels

$I_{D(OFF)}$ = "OFF" leakage of a given FET switch

As an example, if N = 10, $A_D = 0.1\%$, and $I_{D(OFF)} \leq 10$ nA at 85°C for the AH5009. $R1_{(MAX)}$ is:

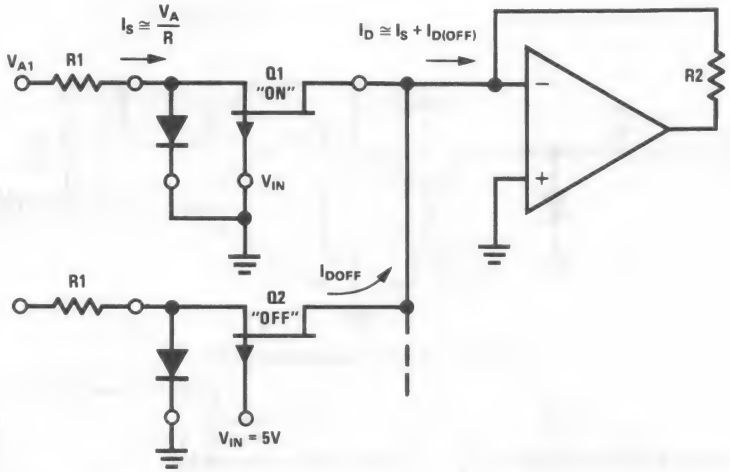
$$R1_{(MAX)} \leq \frac{(1V)(10^{-3})}{(10)(10 \times 10^{-9})} = 10k$$

Selection of R2, of course, depends on the gain desired and for unity gain R1 = R2.

Lastly, the foregoing discussion has ignored resistor tolerances, input bias current and offset voltage of the op amp—all of which should be considered in setting the overall gain accuracy of the circuit.

TTL Compatibility

The AH series can be driven with two different logic voltage swings: the even numbered part types are specified to be driven from standard 5V TTL logic and the odd numbered types from 15V open collector TTL.



TL/H/5659–5

FIGURE 3

Applications Information (Continued)

Standard TTL gates pull-up to about 3.5V (no load). In order to ensure turn-off of the even numbered switches such as AH5010, a pull-up resistor, R_{EXT}, of at least 10 kΩ should be placed between the 5V V_{CC} and the gate output as shown in *Figure 4*.

Likewise, the open-collector, high voltage TTL outputs should use a pull-up resistor as shown in *Figure 5*. In both cases, $t_{(OFF)}$ is improved for lower values of R_{EXT} at the expense of power dissipation in the low state.

Definition of Terms

The terms referred to in the electrical characteristics tables are as defined in *Figure 6*.

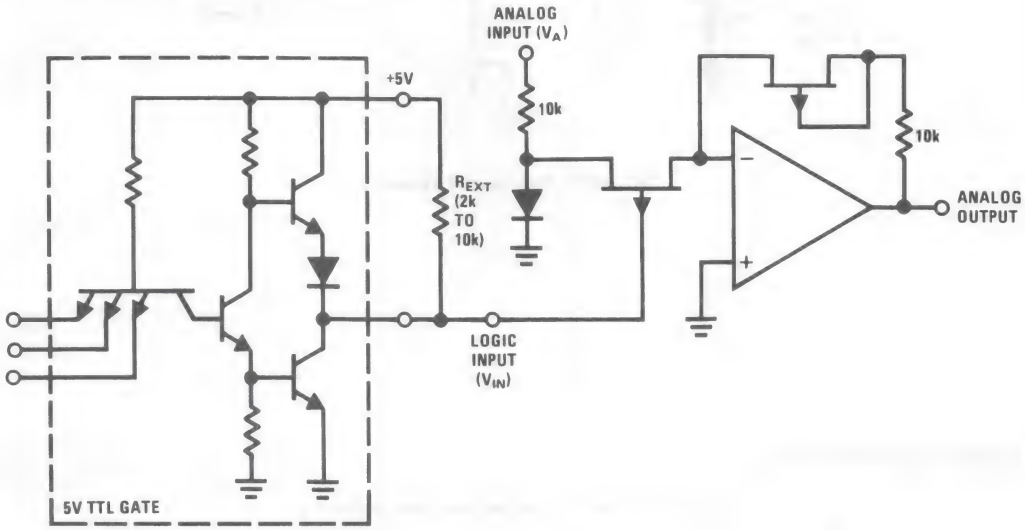
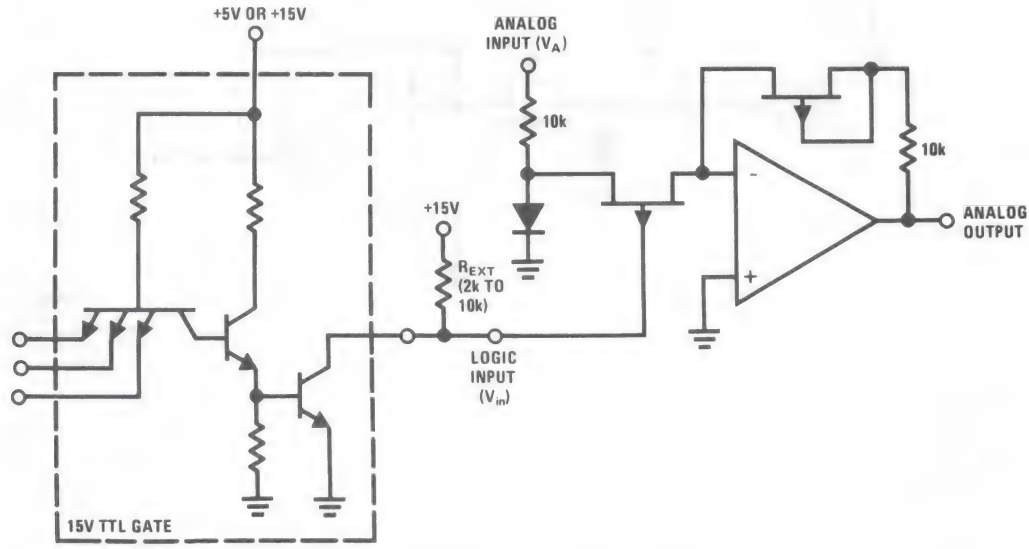


FIGURE 4. Interfacing with +5V TTL



TL/H/5659–6

FIGURE 5. Interfacing with +15V Open Collector TTL

Applications Information (Continued)

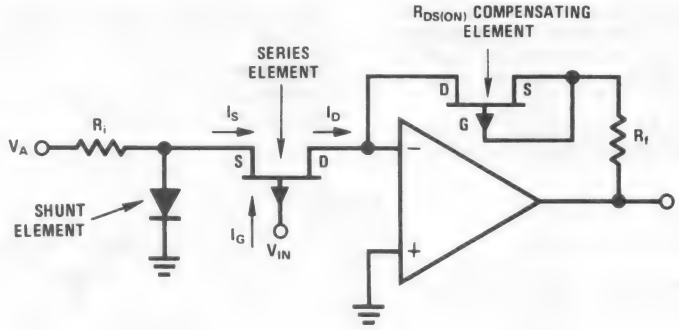
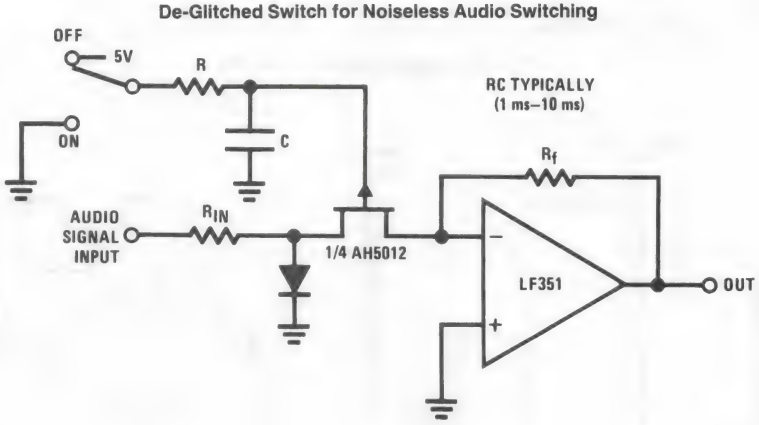


FIGURE 6. Definition of Terms

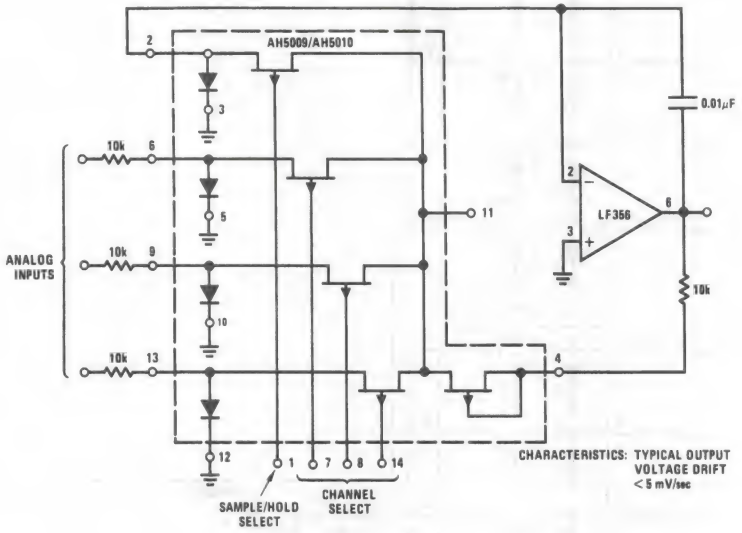
Typical Applications



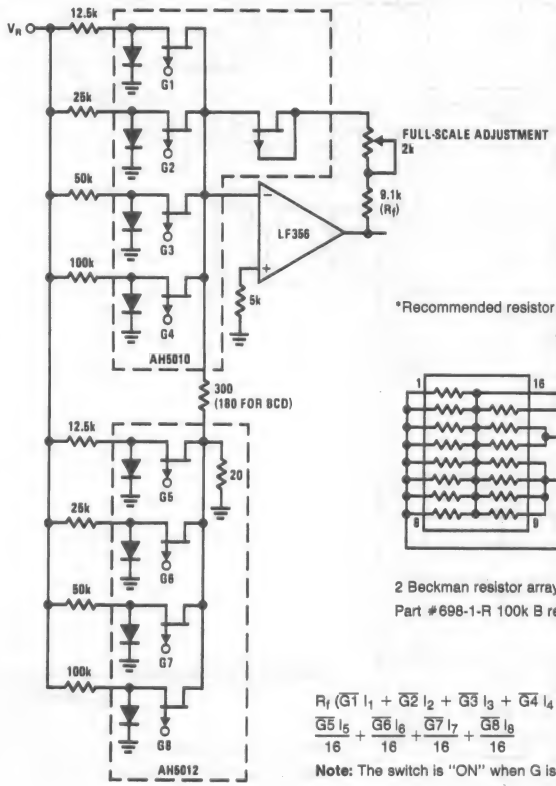
TL/H/5659–7

Typical Applications (Continued)

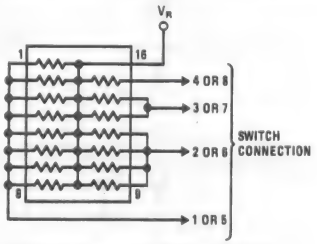
3-Channel Multiplexer with Sample and Hold



8-Bit Binary (BCD) Multiplying D/A Converter*



*Recommended resistor array connection for D/A application



2 Beckman resistor arrays
Part #698-1-R 100k B recommended

$$R_f\left(\frac{\overline{G1}\,I_1 + \overline{G2}\,I_2 + \overline{G3}\,I_3 + \overline{G4}\,I_4 + \overline{G5}\,I_5 + \overline{G6}\,I_6 + \overline{G7}\,I_7 + \overline{G8}\,I_8}{16}\right)$$

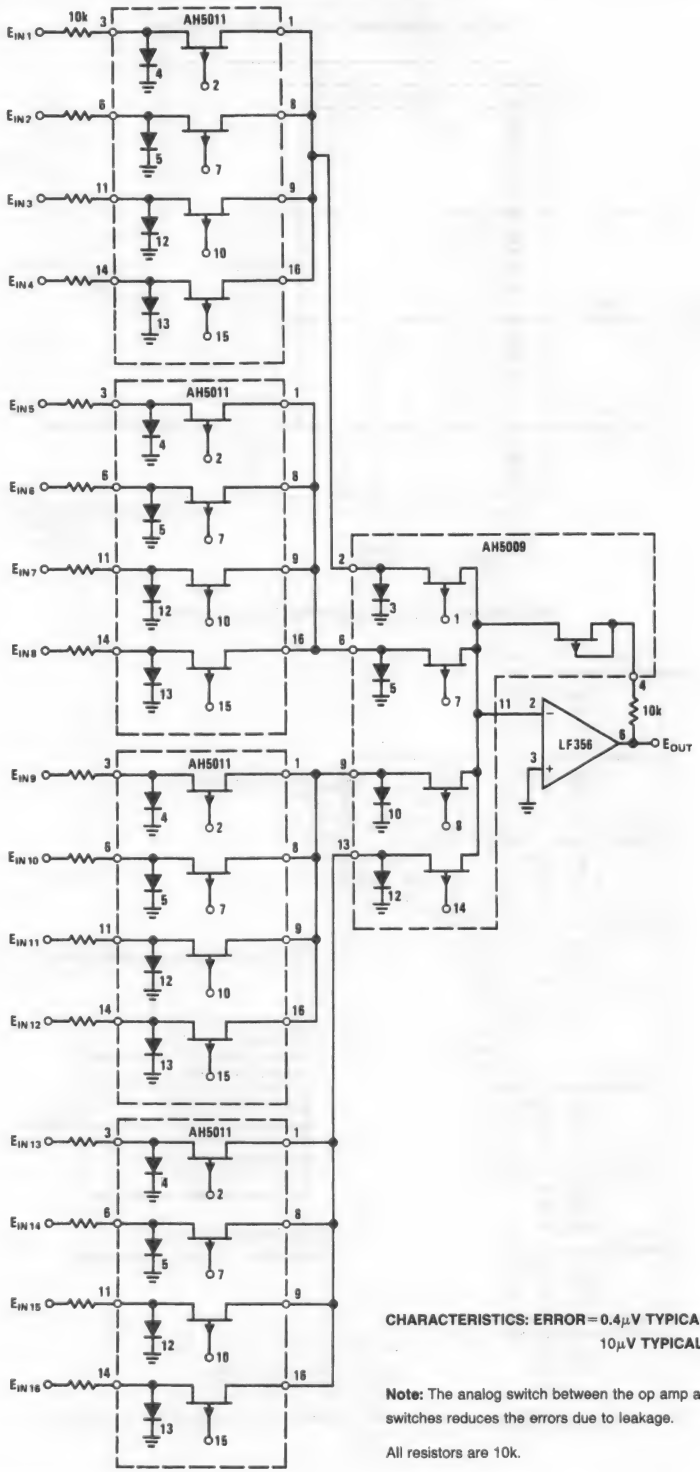
Note: The switch is "ON" when G is at 0V (Logic "0")

$$I = \frac{V_R}{R}$$

TL/H/5659–8

Typical Applications (Continued)

16-Channel Multiplexer



CHARACTERISTICS: ERROR = 0.4μV TYPICAL @ 25°C
10μV TYPICAL @ 70°C

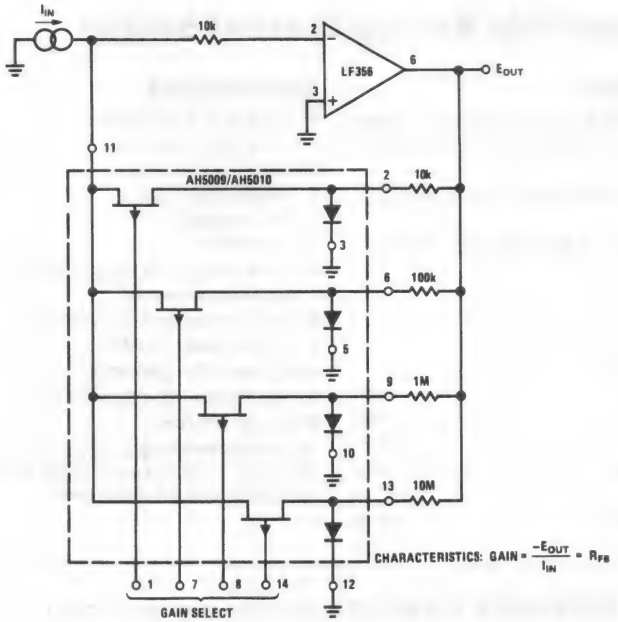
Note: The analog switch between the op amp and the 16 input switches reduces the errors due to leakage.

All resistors are 10k.

TL/H/5659–9

Typical Applications (Continued)

Gain Programmable Amplifier



TL/H/5659–10

National Semiconductor

AH5020C Monolithic Analog Current Switch

General Description

This versatile dual monolithic JFET analog switch economically fulfills a wide variety of multiplexing and analog switching applications.

These switches may be driven directly from standard 5V logic.

The monolithic construction guarantees tight resistance match and track.

Features

■ Interfaces with standard TTL	
■ "ON" resistance match	2Ω
■ Low "ON" resistance	150Ω
■ Very low leakage	50 pA
■ Large analog signal range	±10V peak
■ High switching speed	150 ns
■ Excellent isolation between channels	80 dB at 1 kHz

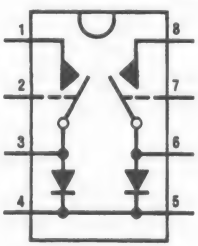
Applications

- A/D and D/A converters
- Micropower converters
- Industrial controllers
- Position controllers
- Data acquisition
- Active filters
- Signal multiplexers/demultiplexers
- Multiple channel AGC
- Quad compressors/expanders
- Choppers/demodulators
- Programmable gain amplifiers
- High impedance voltage buffer
- Sample and hold

For voltage switching applications see LF13201, LF13202, LF13331, LF13332, and LF13333 Analog Switch Family, or the CMOS Analog Switch Family.

Connection and Schematic Diagrams (All switches shown are for logical "1")

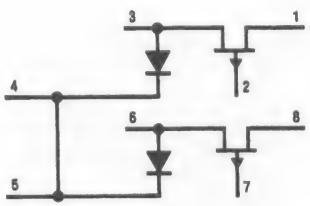
Dual-In-Line Package



TL/H/5166–1

Top View

Order Number AH5020CJ
See NS Package Number J08A



TL/H/5166–2

Note: All diode cathodes are internally connected to the substrate.

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Input Voltage	30V
Positive Analog Signal Voltage	30V
Negative Analog Signal Voltage	−15V
Diode Current	10 mA
Drain Current	30 mA
Power Dissipation	500 mW
Operating Temp. Range	−25°C to +85°C
Storage Temperature Range	−65°C to +150°C
Lead Temp. (Soldering, 10 seconds)	300°C

Electrical Characteristics (Notes 2 and 3)

Symbols	Parameter	Conditions	Typ	Max	Units
I_{GSX}	Input Current "OFF"	V_{GD} = 4.5V, V_{SD} = 0.7V V_{GD} = 11V, V_{SD} = 0.7V T_A = 85°C, V_{GD} = 11V, V_{SD} = 0.7V	0.01 0.01	0.1 0.2 10	nA nA nA
$I_{D(OFF)}$	Leakage Current "OFF"	V_{SD} = 0.7V, V_{GS} = 3.8V T_A = 85°C	0.01	0.2 10	nA nA
$I_{G(ON)}$	Leakage Current "ON"	V_{GD} = 0V, I_S = 1 mA T_A = 85°C	0.08	1 200	nA nA
$I_{G(ON)}$	Leakage Current "ON"	V_{GD} = 0V, I_S = 2 mA T_A = 85°C	0.13	5 10	nA μA
$I_{G(ON)}$	Leakage Current "ON"	V_{GD} = 0V, I_S = −2 mA T_A = 85°C	0.1	10 20	nA μA
$r_{DS(ON)}$	Drain-Source Resistance	V_{GS} = 0.5V, I_S = 2 mA T_A = +85°C	90	150 240	Ω Ω
V_{DIODE}	Forward Diode Drop	I_D = 0.5 mA		0.8	V
$r_{DS(ON)}$	Match	V_{GS} = 0, I_D = 1 mA	2	20	Ω
T_{ON}	Turn "ON" Time	See ac Test Circuit	150	500	ns
T_{OFF}	Turn "OFF" Time	See ac Test Circuit	300	500	ns
CT	Cross Talk	See ac Test Circuit	120		dB

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: Test conditions 25°C unless otherwise noted.

Note 3: "OFF" and "ON" notation refers to the conduction state of the FET switch.

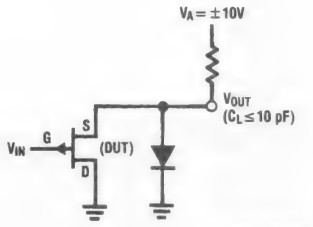
Note 4: Thermal Resistance:

θ_{JA} (Junction to Ambient) ..... N/A

θ_{JC} (Junction to Case) ..... N/A

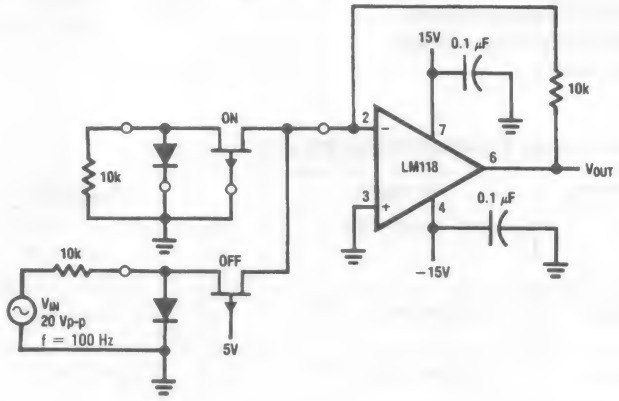
Test Circuits

AC Test Circuit



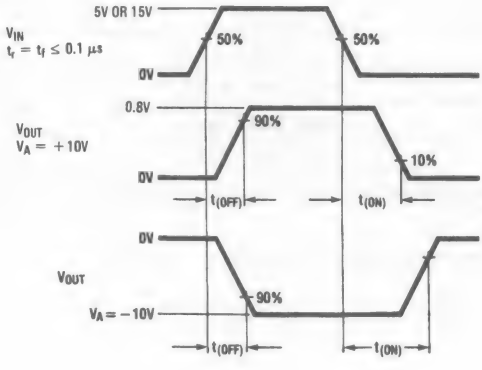
TL/H/5166–4

Cross Talk Test Circuit



TL/H/5166–3

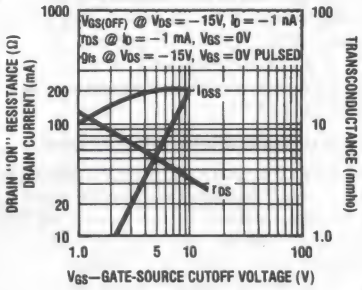
Switching Time Waveforms



TL/H/5166–5

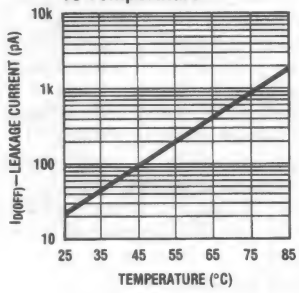
Typical Performance Characteristics

Parameter Interaction



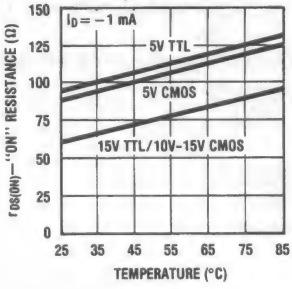
TL/H/5166–6

Leakage Current, $I_{D(OFF)}$ vs Temperature



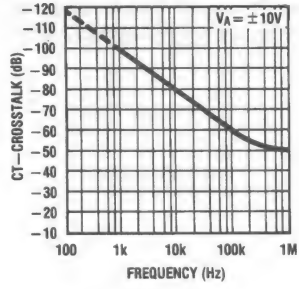
TL/H/5166–7

"ON" Resistance, $r_{DS(ON)}$ vs Temperature



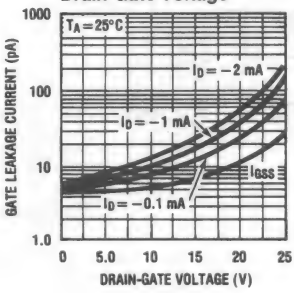
TL/H/5166–8

Cross Talk, CT vs Frequency



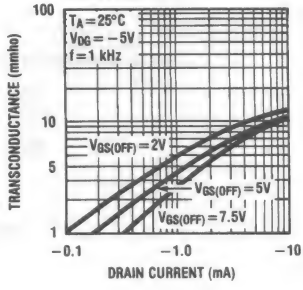
TL/H/5166–9

Leakage Current vs Drain-Gate Voltage



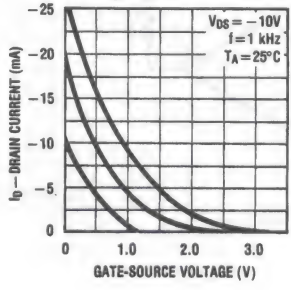
TL/H/5166–10

Transconductance vs Drain Current



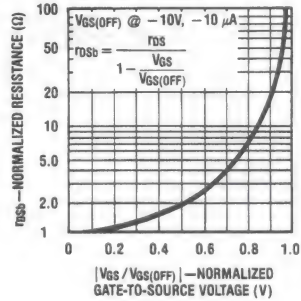
TL/H/5166–11

Drain Current vs Bias Voltage



TL/H/5166–12

Normalized Drain Resistance vs Bias Voltage



TL/H/5166–13

Applications Information

THEORY OF OPERATION

The AH5020 analog switches are primarily intended for operation in current mode switch applications; i.e., the drains of the FET switch are held at or near ground by operating into the summing junction of an operational amplifier. Limiting the drain voltage to under a few hundred millivolts eliminates the need for a special gate driver, allowing the switches to be driven directly by standard TTL.

If only one of the two switches in each package is used to apply an input signal to the input of an op amp, the other switch FET can be placed in the feedback path in order to compensate for the "ON" resistance of the switch FET as shown in *Figure 1*.

The closed-loop gain of *Figure 1* is:

$$A_{VCL} = -\frac{R2 + r_{DS(ON)Q2}}{R1 + r_{DS(ON)Q1}}$$

For R1 = R2, gain accuracy is determined by the $r_{DS(ON)}$ match between Q1 and Q2. Typical match between Q1 and Q2 is 2Ω resulting in a gain accuracy of 0.02% (for R1 = R2 = 10 kΩ).

NOISE IMMUNITY

The switches with the source diodes grounded exhibit improved noise immunity for positive analog signals in the "OFF" state. With V_{IN} = 15V and the V_A = 10V, the source of Q1 is clamped to about 0.7V by the diode (V_{GS} = 14.3V) ensuring that ac signals imposed on the 10V input will not gate the FET "ON".

SELECTION OF GAIN SETTING RESISTORS

Since the AH5020 analog switches are operated in current mode, it is generally advisable to make the signal current as large as possible. However, current through the FET switch tends to forward bias the source to gate junction and the signal shunting diode resulting in leakage through these junctions. As shown in *Figure 2*, $I_{G(ON)}$ represents a finite error in the current reaching the summing junction of the op amp.

Secondly, the $r_{DS(ON)}$ of the FET begins to "round" as I_S approaches I_{DSS}. A practical rule of thumb is to maintain I_S at less than 1/10 of I_{DSS}.

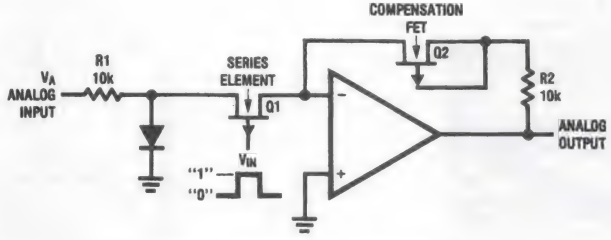
Combining the criteria from the above discussion yields:

$$R1_{(MIN)} \geq \frac{V_{A(MAX)}\, A_D}{I_{G(ON)}} \quad (2a)$$

or:

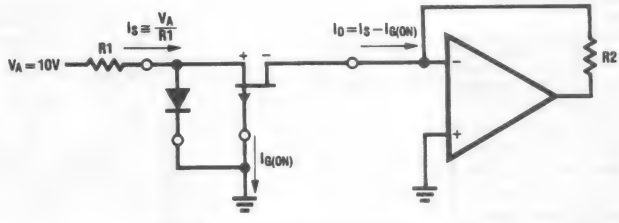
$$\geq \frac{V_{A(MAX)}}{I_{DSS}/10} \quad (2b)$$

whichever is larger.



TL/H/5166–14

FIGURE 1. Use of Compensation FET



TL/H/5166–15

FIGURE 2. On Leakage Current, $I_{G(ON)}$

Applications Information (Continued)

Where $V_{A(MAX)}$ = Peak amplitude of the analog input signal
A_D = Desired accuracy
$I_{G(ON)}$ = Leakage at a given I_S
I_{DSS} = Saturation current of the FET switch = 20 mA

In a typical application, V_A might $= \pm 10V$, $A_D = 0.1\%$, $0°C \leq T_A \leq 85°C$. The criterion of equation (2b) predicts:

$$R1_{(MIN)} \geq \frac{10V}{\frac{20\ mA}{10}} = 5\ k\Omega$$

For R1 = 5k, $I_S \cong 10V/5k$ or 2 mA. The electrical characteristics guarantee an $I_{G(ON)} \leq 1\mu A$ at 85°C for the AH5020. Per the criterion of equation (2a):

$$R1_{(MIN)} \geq \frac{(10V)(10^{-3})}{1 \times 10^{-6}} \geq 10\ k\Omega$$

Since equation (2a) predicts a higher value, the 10k resistor should be used.

The "OFF" condition of the FET also affects gain accuracy. As shown in *Figure 3*, the leakage across Q2, $I_{D(OFF)}$ represents a finite error in the current arriving at the summing junction of the op amp.

Accordingly:

$$R1_{(MAX)} \leq \frac{V_{A(MIN)}\ A_D}{(N)\ I_{D(OFF)}}$$

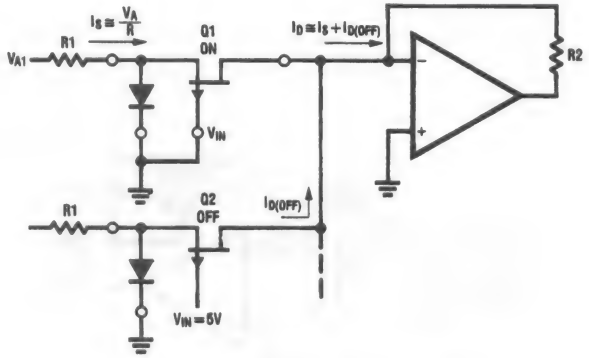
Where $V_{A(MIN)}$ = Minimum value for the analog input signal
A_D = Desired accuracy
N = Number of channels
$I_{D(OFF)}$ = "OFF" leakage of a given FET switch

As an example, if $N = 10$, $A_D = 0.1\%$, and $I_{D(OFF)} \leq 10$ nA at 85°C for the AH5020. $R1_{(MAX)}$ is:

$$R1_{(MAX)} \leq \frac{(1V)(10^{-3})}{(10)(10 \times 10^{-9})} = 10k$$

Selection of R2, of course, depends on the gain desired and for unity gain R1 = R2.

Lastly, the foregoing discussion has ignored resistor tolerances, input bias current and offset voltage of the op amp — all of which should be considered in setting the overall gain accuracy of the circuit.



TL/H/5166–16

FIGURE 3. Off Leakage Current, $I_{D(OFF)}$

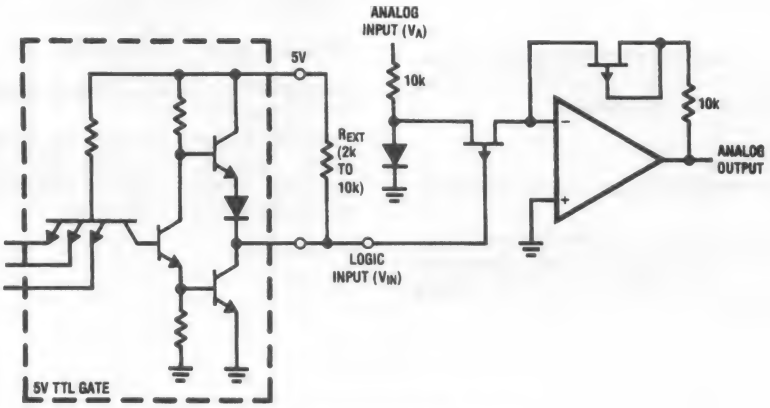
Applications Information (Continued)

TTL COMPATIBILITY

Standard TTL gates pull-up to about 3.5V (no load). In order to ensure turn-off of the AH5020, a pull-up resistor, R_{EXT} of at least 10 kΩ should be placed between the 5V V_{CC} and the gate output as shown in *Figure 4*.

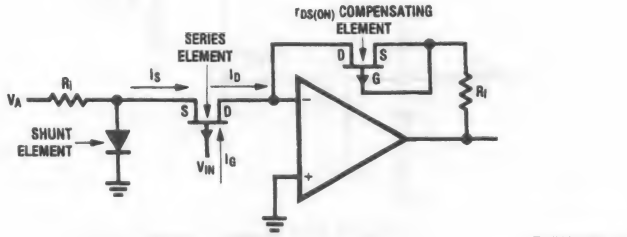
DEFINITION OF TERMS

The terms referred to in the electrical characteristics tables are as defined in *Figure 5*.



TL/H/5166–17

FIGURE 4. Interfacing with +5V TTL

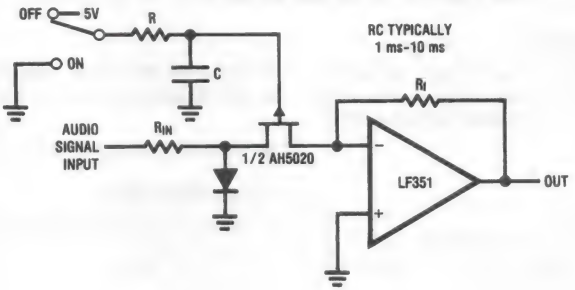


TL/H/5166–18

FIGURE 5. Definition of Terms

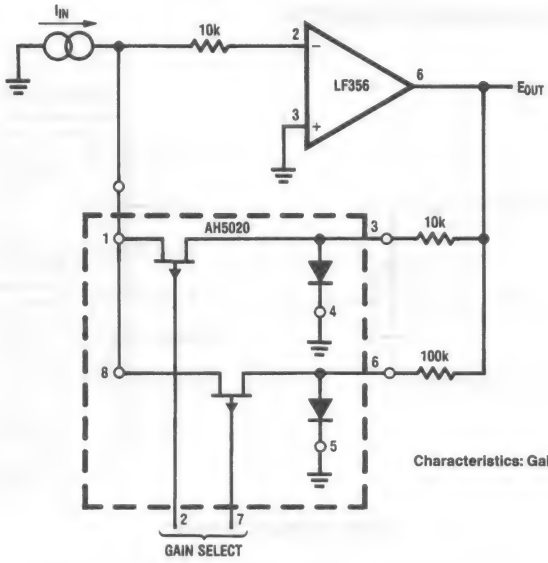
Typical Applications

Deglitched Switch for Noiseless Audio Switching



TL/H/5166–19

Gain Programmable Amplifier



Characteristics: Gain $= \frac{-E_{OUT}}{I_{IN}} = R_{FS}$

TL/H/5166–20

National Semiconductor

CD4016BM/CD4016BC Quad Bilateral Switch

General Description

The CD4016BM/CD4016BC is a quad bilateral switch intended for the transmission or multiplexing of analog or digital signals. It is pin-for-pin compatible with CD4066BM/CD4066BC.

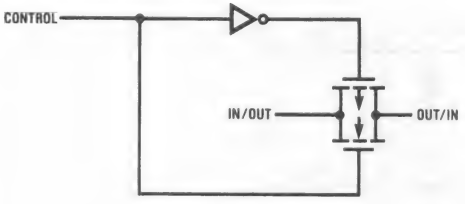
Features

- Wide supply voltage range — 3V to 15V
- Wide range of digital and analog switching — ± 7.5 V_{PEAK}
- "ON" resistance for 15V operation — 400Ω (typ.)
- Matched "ON" resistance over 15V signal input — $\Delta R_{ON} = 10\Omega$ (typ.)
- High degree of linearity — 0.4% distortion (typ.) @ f_{IS} = 1 kHz, $V_{IS} = 5\ V_{p-p}$, $V_{DD} - V_{SS} = 10V$, $R_L = 10\ k\Omega$
- Extremely low "OFF" switch leakage — 0.1 nA (typ.) @ $V_{DD} - V_{SS} = 10V$, $T_A = 25°C$
- Extremely high control input impedance — $10^{12}\Omega$ (typ.)
- Low crosstalk between switches — −50 dB (typ.) @ $f_{IS} = 0.9$ MHz, $R_L = 1\ k\Omega$
- Frequency response, switch "ON" — 40 MHz (typ.)

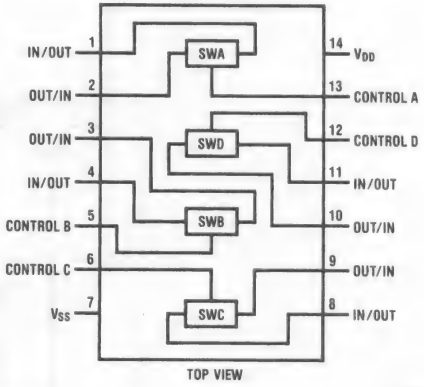
Applications

- Analog signal switching/multiplexing
  - Signal gating
  - Squelch control
  - Chopper
  - Modulator/Demodulator
  - Commutating switch
- Digital signal switching/multiplexing
- CMOS logic implementation
- Analog-to-digital/digital-to-analog conversion
- Digital control of frequency, impedance, phase, and analog-signal gain

Schematic and Connection Diagrams



Dual-In-Line Package



TL/F/5661–1

Order Number CD4016B*

*Please look into Section 8, Appendix D for availability of various package types.

See the CMOS Logic Databook for Complete Specifications

National Semiconductor

CD4051BM/CD4051BC Single 8-Channel Analog Multiplexer/Demultiplexer
CD4052BM/CD4052BC Dual 4-Channel Analog Multiplexer/Demultiplexer
CD4053BM/CD4053BC Triple 2-Channel Analog Multiplexer/Demultiplexer

General Description

These analog multiplexers/demultiplexers are digitally controlled analog switches having low "ON" impedance and very low "OFF" leakage currents. Control of analog signals up to $15V_{p-p}$ can be achieved by digital signal amplitudes of 3–15V. For example, if $V_{DD}=5V$, $V_{SS}=0V$ and $V_{EE}=-5V$, analog signals from $-5V$ to $+5V$ can be controlled by digital inputs of 0–5V. The multiplexer circuits dissipate extremely low quiescent power over the full $V_{DD}-V_{SS}$ and $V_{DD}-V_{EE}$ supply voltage ranges, independent of the logic state of the control signals. When a logical "1" is present at the inhibit input terminal all channels are "OFF".

CD4051BM/CD4051BC is a single 8-channel multiplexer having three binary control inputs. A, B, and C, and an inhibit input. The three binary signals select 1 of 8 channels to be turned "ON" and connect the input to the output.

CD4052BM/CD4052BC is a differential 4-channel multiplexer having two binary control inputs, A and B, and an inhibit input. The two binary input signals select 1 or 4 pairs of channels to be turned on and connect the differential analog inputs to the differential outputs.

CD4053BM/CD4053BC is a triple 2-channel multiplexer having three separate digital control inputs, A, B, and C, and an inhibit input. Each control input selects one of a pair of channels which are connected in a single-pole double-throw configuration.

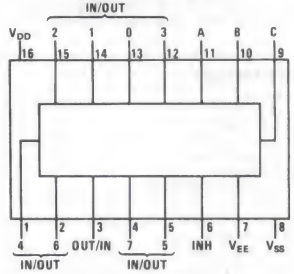
Features

- Wide range of digital and analog signal levels: digital 3–15V, analog to $15V_{p-p}$
- Low "ON" resistance: 80Ω (typ.) over entire $15V_{p-p}$ signal-input range for $V_{DD}-V_{EE}=15V$
- High "OFF" resistance: channel leakage of ±10 pA (typ.) at $V_{DD}-V_{EE}=10V$
- Logic level conversion for digital addressing signals of 3–15V ($V_{DD}-V_{SS}=3-15V$) to switch analog signals to 15 V_{p-p} ($V_{DD}-V_{EE}=15V$)
- Matched switch characteristics: $\Delta R_{ON}=5\Omega$ (typ.) for $V_{DD}-V_{EE}=15V$
- Very low quiescent power dissipation under all digital-control input and supply conditions: 1 μW (typ.) at $V_{DD}-V_{SS}=V_{DD}-V_{EE}=10V$
- Binary address decoding on chip

Connection Diagrams

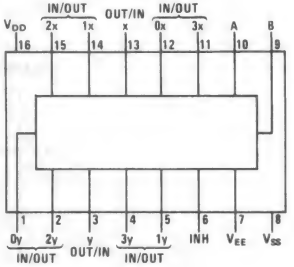
Dual-In-Line Packages

CD4051BM/CD4051BC



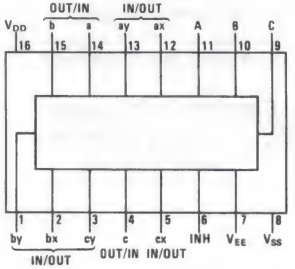
TOP VIEW

CD4052BM/CD4052BC



TOP VIEW

CD4053BM/CD4053BC



TOP VIEW TL/F/5662–1

Order Number CD4051B*, CD4052B*, or CD4053B*

*Please look into Section 8, Appendix D for availability of various package types.

See the CMOS Logic Databook for Complete Specifications

National Semiconductor

CD4066BM/CD4066BC Quad Bilateral Switch

General Description

The CD4066BM/CD4066BC is a quad bilateral switch intended for the transmission or multiplexing of analog or digital signals. It is pin-for-pin compatible with CD4016BM/CD4016BC, but has a much lower "ON" resistance, and "ON" resistance is relatively constant over the input-signal range.

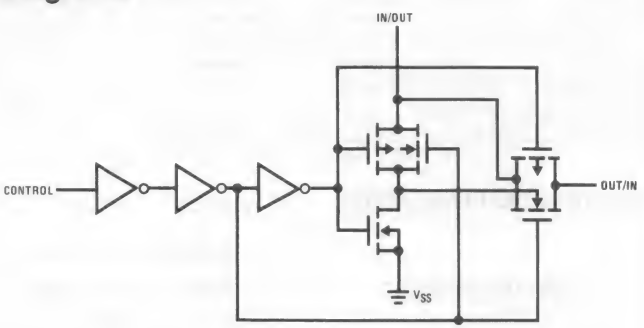
Features

- Wide supply voltage range — 3V to 15V
- High noise immunity — 0.45 V_{DD} (typ.)
- Wide range of digital and analog switching — ±7.5 V_{PEAK}
- "ON" resistance for 15V operation — 80Ω
- Matched "ON" resistance over 15V signal input — ΔR_{ON} = 5Ω (typ.)
- "ON" resistance flat over peak-to-peak signal range
- High "ON"/"OFF" output voltage ratio — 65 dB (typ.) @ f_{is} = 10 kHz, R_L = 10 kΩ
- High degree linearity — 0.1% distortion (typ.)
  High degree linearity — @ f_{is} = 1 kHz, V_{is} = 5V_{p-p},
  High degree linearity — $V_{DD} - V_{SS}$ = 10V, R_L = 10 kΩ
- Extremely low "OFF" switch leakage — 0.1 nA (typ.) @ $V_{DD} - V_{SS}$ = 10V, T_A = 25°C
- Extremely high control input impedance — 10^{12}Ω (typ.)
- Low crosstalk between switches — −50 dB (typ.) @ f_{is} = 0.9 MHz, R_L = 1 kΩ
- Frequency response, switch "ON" — 40 MHz (typ.)

Applications

- Analog signal switching/multiplexing
  - Signal gating
  - Squelch control
  - Chopper
  - Modulator/Demodulator
  - Commutating switch
- Digital signal switching/multiplexing
- CMOS logic implementation
- Analog-to-digital/digital-to-analog conversion
- Digital control of frequency, impedance, phase, and analog-signal-gain

Schematic and Connection Diagrams

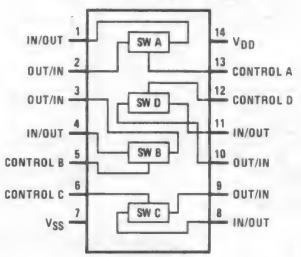


Order Number CD4066B*

*Please look into Section 8, Appendix D for availability of various package types.

See the CMOS Logic Databook for Complete Specifications

Dual-In-Line Package



TL/F/5665–1

Top View

National Semiconductor

CD4529BM/CD4529BC Dual 4-Channel or Single 8-Channel Analog Data Selector

General Description

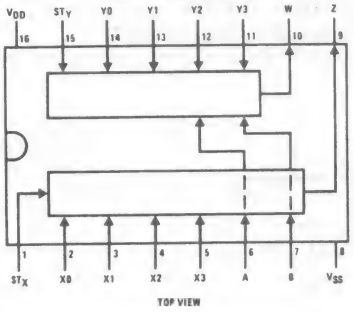
The CD4529B is a dual 4-channel or a single 8-channel analog data selector, implemented with complementary MOS (CMOS) circuits constructed with N- and P-channel enhancement mode transistors. Dual 4-channel or 8-channel mode operation is selected by proper input coding, with outputs Z and W tied together for the single 8-bit mode. The device is suitable for digital as well as analog applications, including various 1-of-4 and 1-of-8 data selector functions. Since the device is analog and bidirectional, it can also be used for dual binary to 1-of-4 or single 1-of-8 decoder applications.

Features

- Wide supply voltage range — 3.0V to 15V
- High noise immunity — 0.45 V_{DD} (typ.)
- Low quiescent power dissipation — 0.005 μW/package (typ.)@5.0 V_{DC}
- 10 MHz frequency operation (typ.)
- Data paths are bidirectional
- Linear ON resistance [120Ω (typ.)@15V]
- TRI-STATE® outputs (high impedance disable strobe)
- Plug-in replacement for MC14529B

Connection Diagram

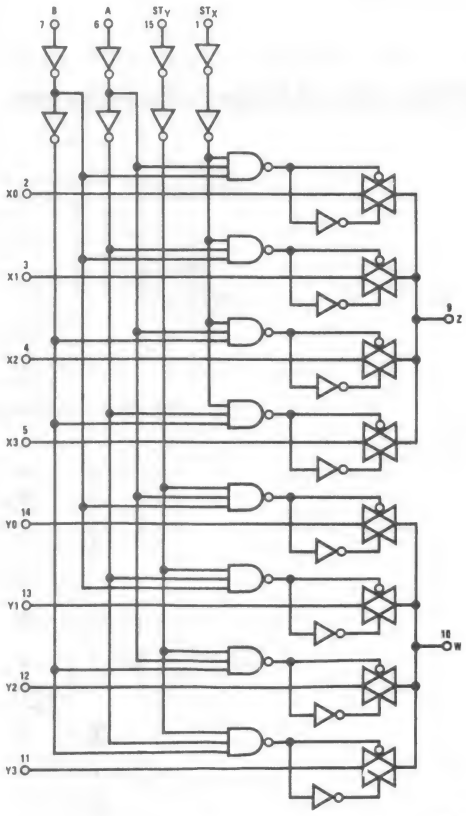
Dual-In-Line Package



Order Number CD4529B*

*Please look into Section 8, Appendix D for availability of various package types.

Logic Diagram



TL/F/5999–1

Truth Table

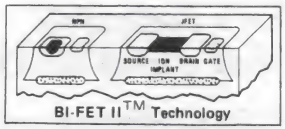
ST_X	ST_Y	B	A	Z	W	
1	1	0	0	X0	Y0	Dual 4-Channel Mode 2 Outputs
1	1	0	1	X1	Y1	
1	1	1	0	X2	Y2	
1	1	1	1	X3	Y3	
1	0	0	0	X0		Single 8-Channel Mode 1 Output (Z and W tied together)
1	0	0	1	X1		
1	0	1	0	X2		
1	0	1	1	X3		
0	1	0	0	Y0		
0	1	0	1	Y1		
0	1	1	0	Y2		
0	1	1	1	Y3		
0	0	X	X	High Impedance (TRI-STATE)		

X = Don't care

See the CMOS Logic Databook for Complete Specifications

National Semiconductor

Quad SPST JFET Analog Switches



LF11331, LF13331 4 Normally Open Switches with Disable
LF11332, LF13332 4 Normally Closed Switches with Disable
LF11333, LF13333 2 Normally Closed Switches and 2 Normally Open Switches with Disable
LF11201, LF13201 4 Normally Closed Switches
LF11202, LF13202 4 Normally Open Switches

General Description

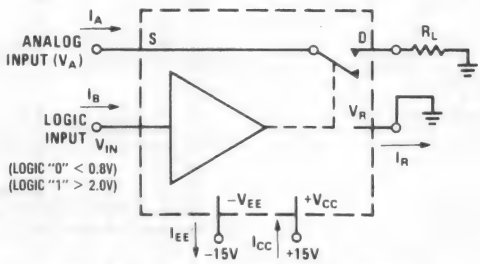
These devices are a monolithic combination of bipolar and JFET technology producing the industry's first one chip quad JFET switch. A unique circuit technique is employed to maintain a constant resistance over the analog voltage range of ±10V. The input is designed to operate from minimum TTL levels, and switch operation also ensures a break-before-make action.

These devices operate from ±15V supplies and swing a ±10V analog signal. The JFET switches are designed for applications where a dc to medium frequency analog signal needs to be controlled.

Features

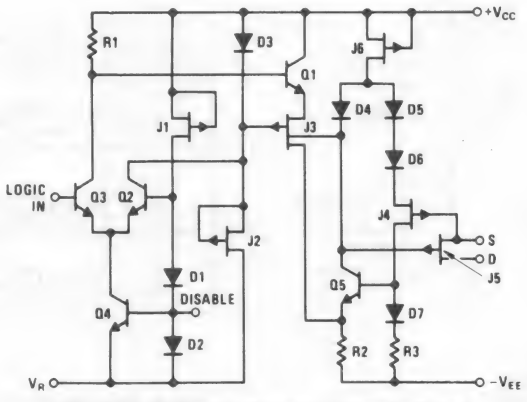
- Analog signals are not loaded
- Constant "ON" resistance for signals up to ±10V and 100 kHz
- Pin compatible with CMOS switches with the advantage of blow out free handling
- Small signal analog signals to 50 MHz
- Break-before-make action $t_{OFF} < t_{ON}$
- High open switch isolation at 1.0 MHz −50 dB
- Low leakage in "OFF" state <1.0 nA
- TTL, DTL, RTL compatibility
- Single disable pin opens all switches in package on LF11331, LF11332, LF11333
- LF11201 is pin compatible with DG201

Test Circuit and Schematic Diagram



TL/H/5667–2

FIGURE 1. Typical Circuit for One Switch



TL/H/5667–12

FIGURE 2. Schematic Diagram (Normally Open)

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications. (Note 1)

Supply Voltage ($V_{CC}-V_{EE}$)	36V
Reference Voltage	$V_{EE} \le V_R \le V_{CC}$
Logic Input Voltage	$V_R-4.0V \le V_{IN} \le V_R+6.0V$
Analog Voltage	$V_{EE} \le V_A \le V_{CC}+6V$; $V_A \le V_{EE}+36V$
Analog Current	$\lvert I_A \rvert < 20$ mA

Power Dissipation (Note 2)	
Molded DIP (N Suffix)	500 mW
Cavity DIP (D Suffix)	900 mW
Operating Temperature Range	
LF11201, 2 and LF11331, 2, 3	−55°C to +125°C
LF13201, 2 and LF13331, 2, 3	0°C to +70°C
Storage Temperature	−65°C to +150°C
Soldering Information	
N and D Package (10 sec.)	300°C
SO Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

Electrical Characteristics (Note 3)

Symbol	Parameter	Conditions		LF11331/2/3 LF11201/2			LF13331/2/3 LF13201/2			Units
				Min	Typ	Max	Min	Typ	Max	
R_{ON}	"ON" Resistance	$V_A=0$, $I_D=1$ mA	$T_A=25°C$		150	200		150	250	Ω
					200	300		200	350	Ω
R_{ON} Match	"ON" Resistance Matching		$T_A=25°C$		5	20		10	50	Ω
V_A	Analog Range			±10	±11		±10	±11		V
$I_{S(ON)}$ + $I_{D(ON)}$	Leakage Current in "ON" Condition	Switch "ON," $V_S=V_D=\pm10V$	$T_A=25°C$		0.3	5		0.3	10	nA
					3	100		3	30	nA
$I_{S(OFF)}$	Source Current in "OFF" Condition	Switch "OFF," $V_S=+10V$, $V_D=-10V$	$T_A=25°C$		0.4	5		0.4	10	nA
					3	100		3	30	nA
$I_{D(OFF)}$	Drain Current in "OFF" Condition	Switch "OFF," $V_S=+10V$, $V_D=-10V$	$T_A=25°C$		0.1	5		0.1	10	nA
					3	100		3	30	nA
V_{INH}	Logical "1" Input Voltage			2.0			2.0			V
V_{INL}	Logical "0" Input Voltage					0.8			0.8	V
I_{INH}	Logical "1" Input Current	$V_{IN}=5V$	$T_A=25°C$		3.6	10		3.6	40	μA
						25			100	μA
I_{INL}	Logical "0" Input Current	$V_{IN}=0.8$	$T_A=25°C$			0.1			0.1	μA
						1			1	μA
t_{ON}	Delay Time "ON"	$V_S=\pm10V$, *(Figure 3)*	$T_A=25°C$		500			500		ns
t_{OFF}	Delay Time "OFF"	$V_S=\pm10V$, *(Figure 3)*	$T_A=25°C$		90			90		ns
$t_{ON}-t_{OFF}$	Break-Before-Make	$V_S=\pm10V$, *(Figure 3)*	$T_A=25°C$		80			80		ns
$C_{S(OFF)}$	Source Capacitance	Switch "OFF," $V_S=\pm10V$	$T_A=25°C$		4.0			4.0		pF
$C_{D(OFF)}$	Drain Capacitance	Switch "OFF," $V_D=\pm10V$	$T_A=25°C$		3.0			3.0		pF
$C_{S(ON)}$ + $C_{D(ON)}$	Active Source and Drain Capacitance	Switch "ON," $V_S=V_D=0V$	$T_A=25°C$		5.0			5.0		pF
$I_{SO(OFF)}$	"OFF" Isolation	*(Figure 4)*, (Note 4)	$T_A=25°C$		−50			−50		dB
CT	Crosstalk	*(Figure 4)*, (Note 4)	$T_A=25°C$		−65			−65		dB
SR	Analog Slew Rate	(Note 5)	$T_A=25°C$		50			50		V/μs
I_{DIS}	Disable Current	*(Figure 5)*, (Note 6)	$T_A=25°C$		0.4	1.0		0.6	1.5	mA
					0.6	1.5		0.9	2.3	mA
I_{EE}	Negative Supply Current	All Switches "OFF," $V_S=\pm10V$	$T_A=25°C$		3.0	5.0		4.3	7.0	mA
					4.2	7.5		6.0	10.5	mA
I_R	Reference Supply Current	All Switches "OFF," $V_S=\pm10V$	$T_A=25°C$		2.0	4.0		2.7	5.0	mA
					2.8	6.0		3.8	7.5	mA
I_{CC}	Positive Supply Current	All Switches "OFF," $V_S=\pm10V$	$T_A=25°C$		4.5	6.0		7.0	9.0	mA
					6.3	9.0		9.8	13.5	mA

Note 1: Refer to RETSF11201X, RETSF11331X, RETSF11332X and RETSF11333X for military specifications.

Note 2: For operating at high temperature the molded DIP products must be derated based on a +100°C maximum junction temperature and a thermal resistance of +150°C/W, devices in the cavity DIP are based on a +150°C maximum junction temperature and are derated at ±100°C/W.

Note 3: Unless otherwise specified, $V_{CC}=+15V$, $V_{EE}=-15V$, $V_R=0V$, and limits apply for $-55°C \le T_A \le +125°C$ for the LF11331/2/3 and the LF11201/2, $-25°C \le T_A \le +85°C$ for the LF13331/2/3 and the LF13201/2.

Note 4: These parameters are limited by the pin to pin capacitance of the package.

Note 5: This is the analog signal slew rate above which the signal is distorted as a result of finite internal slew rates.

Note 6: All switches in the device are turned "OFF" by saturating a transistor at the disable node as shown in *Figure 5*. The delay time will be approximately equal to the t_{ON} or t_{OFF} plus the delay introduced by the external transistor.

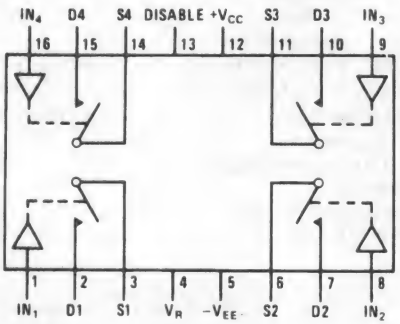
Note 7: This graph indicates the analog current at which 1% of the analog current is lost when the drain is positive with respect to the source.

Note 8: θ_{JA} (Typical) Thermal Resistance

Molded DIP (N)	85°C/W
Cavity DIP (D)	100°C/W
Small Outline (M)	105°C/W

Connection Diagrams (Top View for SO and Dual-In-Line Packages) (All Switches Shown are For Logical "0")

LF11331/LF13331



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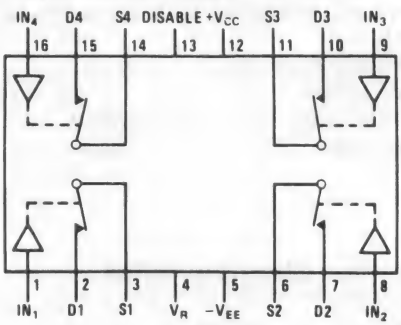
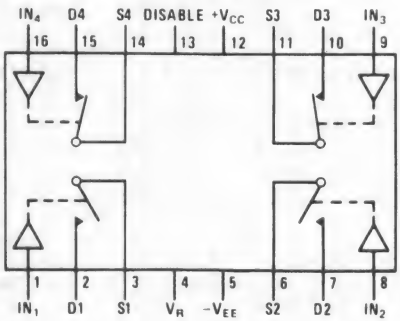
LF11332/LF13332



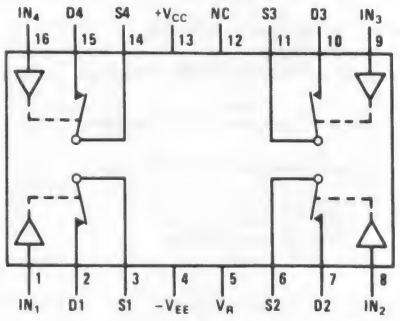
TL/H/5667–13

LF11333/LF13333



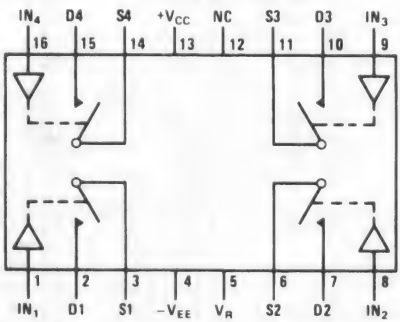
TL/H/5667–14

LF11201/LF13201



TL/H/5667–15

LF11202/LF13202



TL/H/5667–16

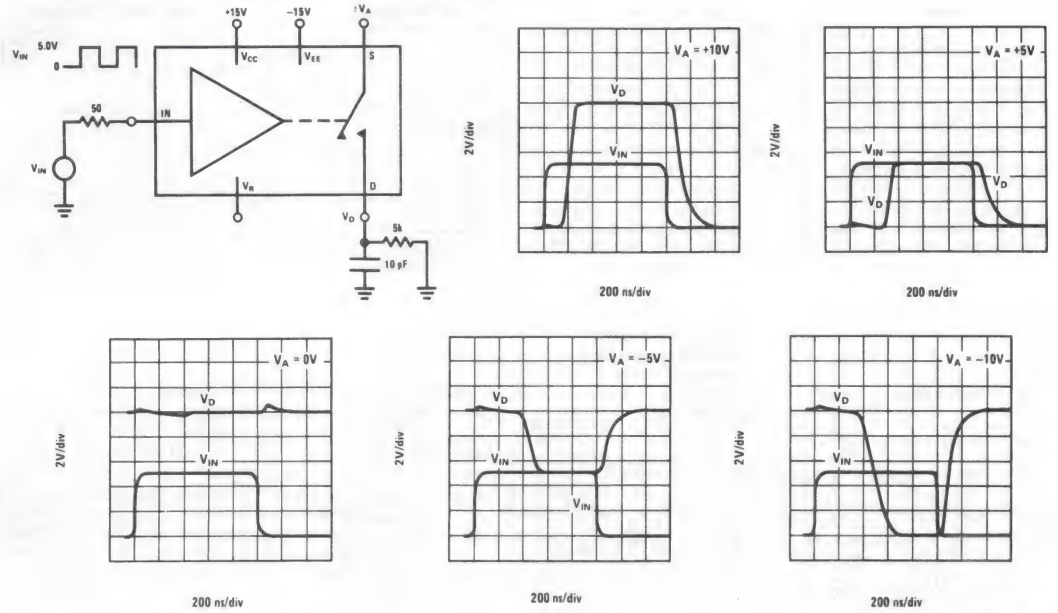
Order Number LF13201D, LF11201D, LF13202D, LF11202D, LF13331D, LF11331D, LF13332D, LF11332D, LF13333D or LF11333D
See NS Package Number D16C

Order Number LF13201M, LF13202M, LF13331M, LF13332M or LF13333M
See NS Package Number M16A

Order Number LF13201N, LF13202N, LF13331N, LF13332N or LF13333N
See NS Package Number N16A

Test Circuit and Typical Performance Curves

Delay Time, Rise Time, Settling Time, and Switching Transients



TL/H/5667–3

Additional Test Circuits

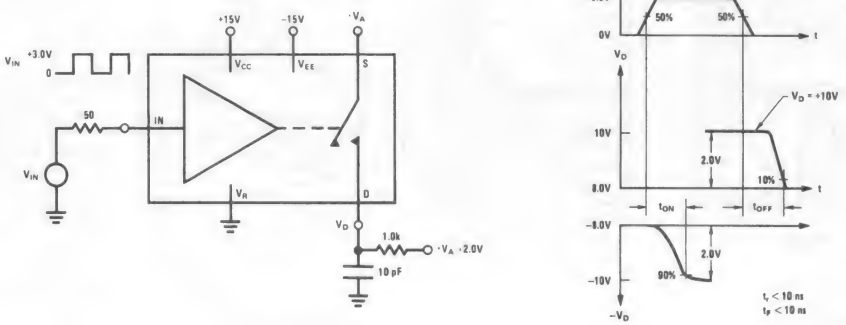
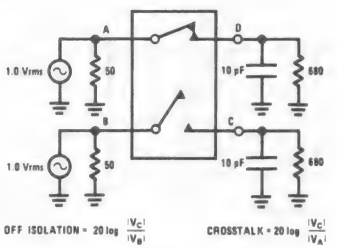


FIGURE 3. t_{ON}, t_{OFF} Test Circuit and Waveforms for a Normally Open Switch

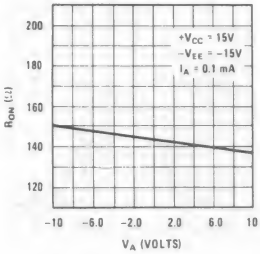


TL/H/5667–4

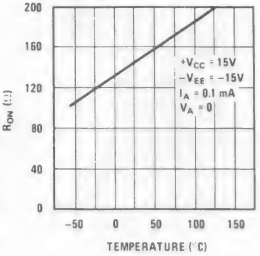
FIGURE 4. "OFF" Isolation, Crosstalk, Small Signal Response

Typical Performance Characteristics

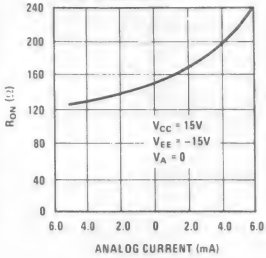
"ON" Resistance



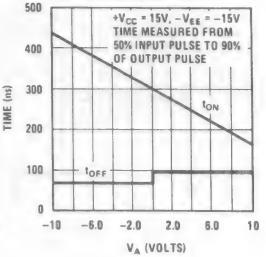
"ON" Resistance



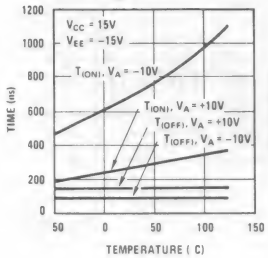
"ON" Resistance



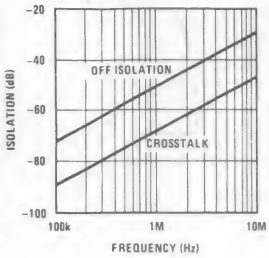
Break-Before-Make Action



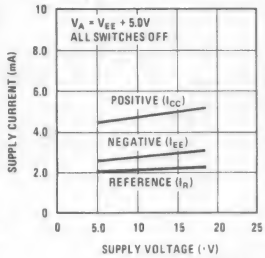
Switching Times



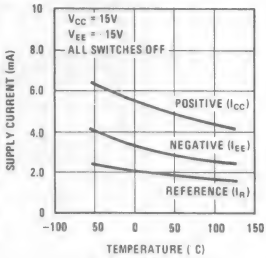
Crosstalk and "OFF" Isolation vs Frequency Using Test Circuit of *Figure 5*



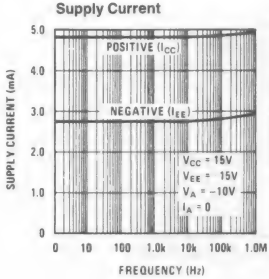
Supply Current



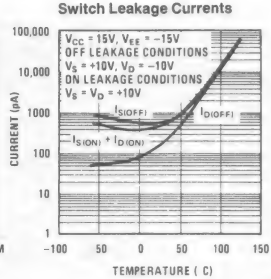
Supply Current



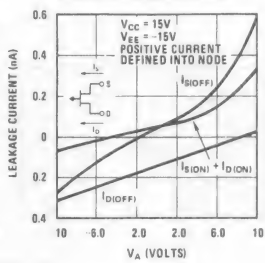
Supply Current



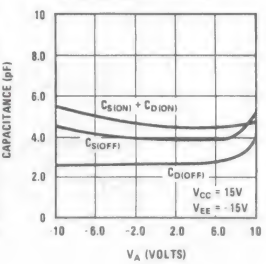
Switch Leakage Currents



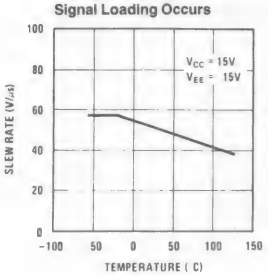
Switch Leakage Current



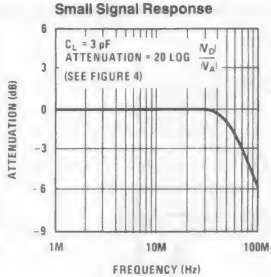
Switch Capacitances



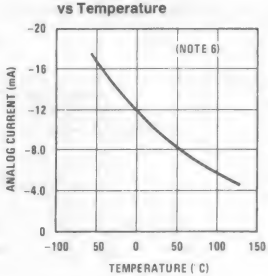
Slew Rate of Analog Voltage Above Which Signal Loading Occurs



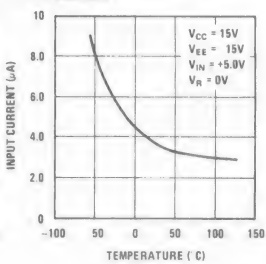
Small Signal Response



Maximum Accurate Analog Current vs Temperature



Logical "1" Input Bias Current



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Application Hints

GENERAL INFORMATION

These devices are monolithic quad JFET analog switches with "ON" resistances which are essentially independent of analog voltage or analog current. The leakage currents are typically less than 1 nA at 25°C in both the "OFF"and "ON" switch states and introduce negligible errors in most applications. Each switch is controlled by minimum TTL logic levels at its input and is designed to turn "OFF" faster than it will turn "ON." This prevents two analog sources from being transiently connected together during switching. The switches were designed for applications which require break-before-make action, no analog current loss, medium speed switching times and moderate analog currents.

Because these analog switches are JFET rather than CMOS, they do not require special handling.

LOGIC INPUTS

The logic input (IN), of each switch, is referenced to two forward diode drops (1.4V at 25°C) from the reference supply (V_R) which makes it compatible with DTL, RTL, and TTL logic families. For normal operation, the logic "0" voltage can range from 0.8V to -4.0V with respect to V_R and the logic "1" voltage can range from 2.0V to 6.0V with respect to V_R, provided V_{IN} is not greater than ($V_{CC}-2.5$V). If the input voltage is greater than ($V_{CC}-2.5$V), the input current will increase. If the input voltage exceeds 6.0V or -4.0V with respect to V_R, a resistor in series with the input should be used to limit the input current to less than 100μA.

ANALOG VOLTAGE AND CURRENT

Analog Voltage

Each switch has a constant "ON" resistance (R_{ON}) for analog voltages from ($V_{EE}+5$V) to ($V_{CC}-5$V). For analog voltages greater than ($V_{CC}-5$V), the switch will remain ON independent of the logic input voltage. For analog voltages less than ($V_{EE}+5$V), the ON resistance of the switch will increase. Although the switch will not operate normally when the analog voltage is out of the previously mentioned range, the source voltage can go to either ($V_{EE}+36$V) or ($V_{CC}+6$V), whichever is more positive, and can go as negative as V_{EE} without destruction. The drain (D) voltage can also go to either ($V_{EE}+36$V) or ($V_{CC}+6$V), whichever is more positive, and can go as negative as ($V_{CC}-36$V) without destruction.

Analog Current

With the source (S) positive with respect to the drain (D), the R_{ON} is constant for low analog currents, but will increase at higher currents (>5 mA) when the FET enters the saturation region. However, if the drain is positive with respect to the source and a small analog current loss at high analog currents (Note 6) is tolerable, a low R_{ON} can be maintained for analog currents greater than 5 mA at 25°C.

LEAKAGE CURRENTS

The drain and source leakage currents, in both the ON and the OFF states of each switch, are typically less than 1 nA at 25°C and less than 100 nA at 125°C. As shown in the typical curves, these leakage currents are Dependent on power supply voltages, analog voltage, analog current and the source to drain voltage.

DELAY TIMES

The delay time OFF (t_{OFF}) is essentially independent of both the analog voltage and temperature. The delay time ON (t_{ON}) will decrease as either ($V_{CC}-V_A$) decreases or the temperature decreases.

POWER SUPPLIES

The voltage between the positive supply (V_{CC}) and either the negative supply (V_{EE}) or the reference supply (V_R) can be as much as 36V. To accommodate variations in input logic reference voltages, V_R can range from V_{EE} to ($V_{CC}-4.5$V). Care should be taken to ensure that the power supply leads for the device never become reversed in polarity or that the device is never inadvertantly installed backwards in a test socket. If one of these conditions occurs, the supplies would zener an internal diode to an unlimited current; and result in a destroyed device.

SWITCHING TRANSIENTS

When a switch is turned OFF or ON, transients will appear at the load due to the internal transient voltage at the gate of the switch JFET being coupled to the drain and source by the junction capacitances of the JFET. The magnitude of these transients is dependent on the load. A lower value R_L produces a lower transient voltage. A negative transient occurs during the delay time ON, while a positive transient occurs during the delay time OFF. These transients are relatively small when compared to faster switch families.

DISABLE NODE

This node can be used, as shown in *Figure 5*, to turn all the switches in the unit off independent of logic inputs. Normally, the node floats freely at an internal diode drop ($\approx$0.7V) above V_R. When the external transistor in *Figure 5* is saturated, the node is pulled very close to V_R and the unit is disabled. Typically, the current from the node will be less than 1 mA. This feature is not available on the LF11201 or LF11202 series.

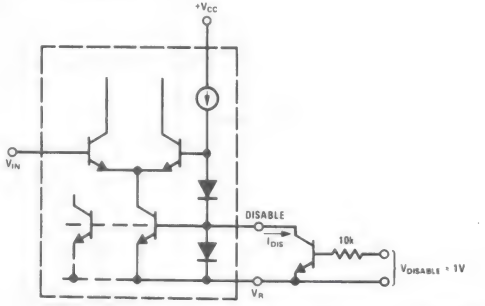

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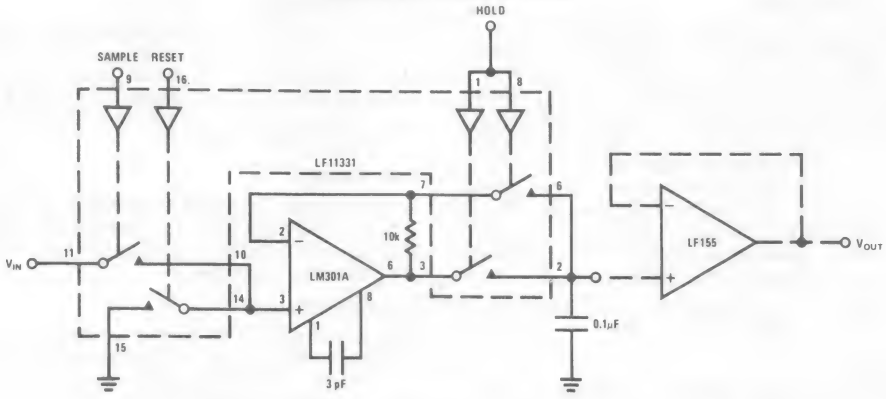
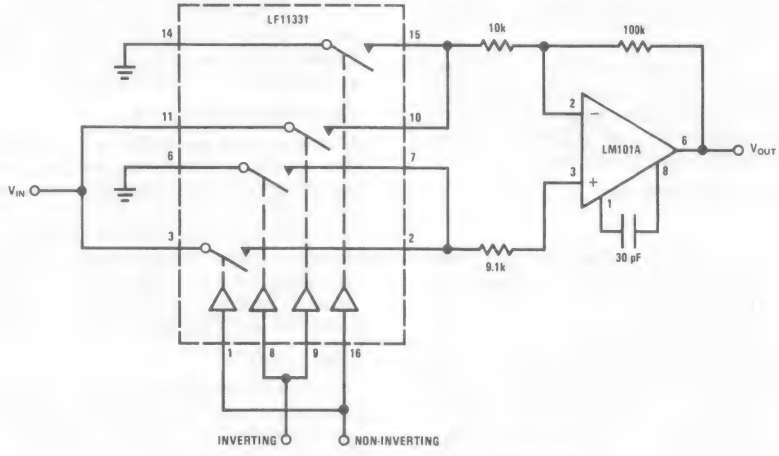
FIGURE 5. Disable Function

Typical Applications

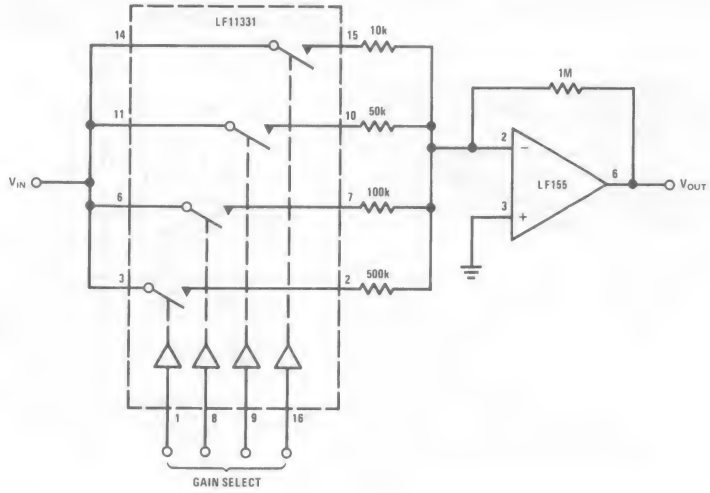
Sample and Hold with Reset



Programmable Inverting Non-Inverting Operational Amplifier



Programmable Gain Operational Amplifier



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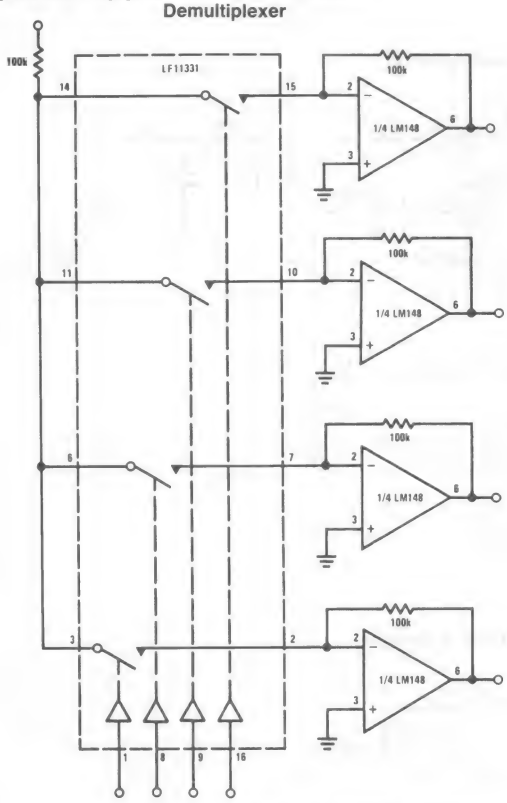
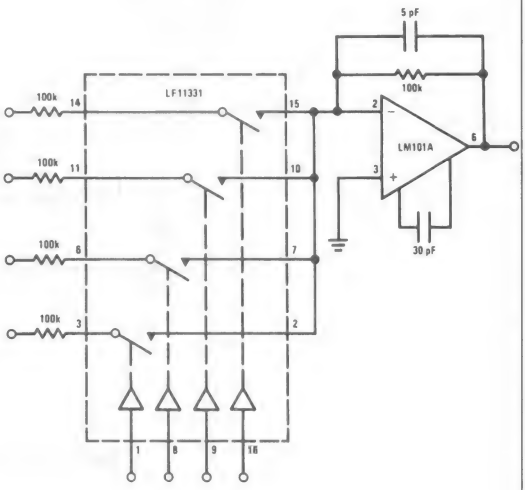
Typical Applications (Continued)

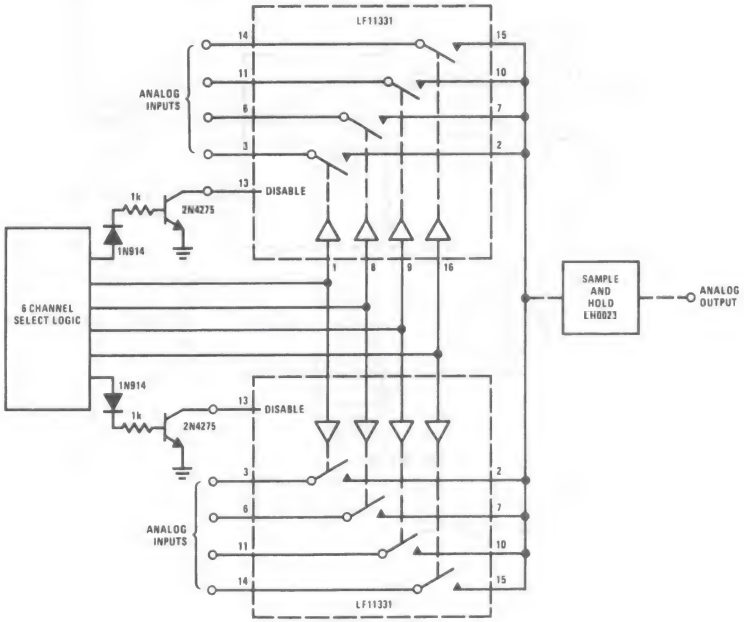
Demultiplexer



Multiplexer/Mixer



8-Channel Analog Commutator with 6-Channel Select Logic



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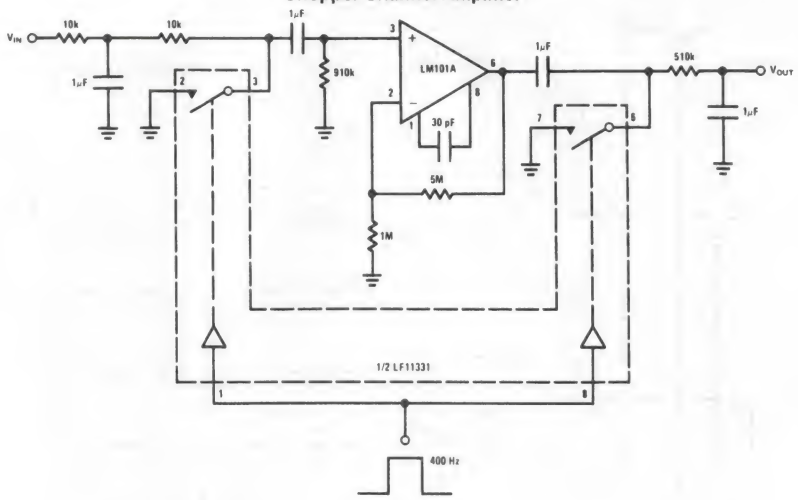
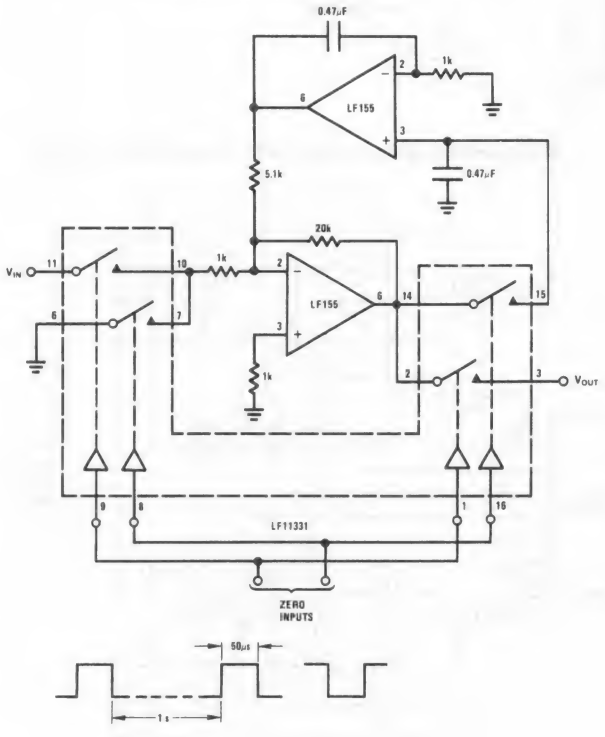
Typical Applications (Continued)

Chopper Channel Amplifier



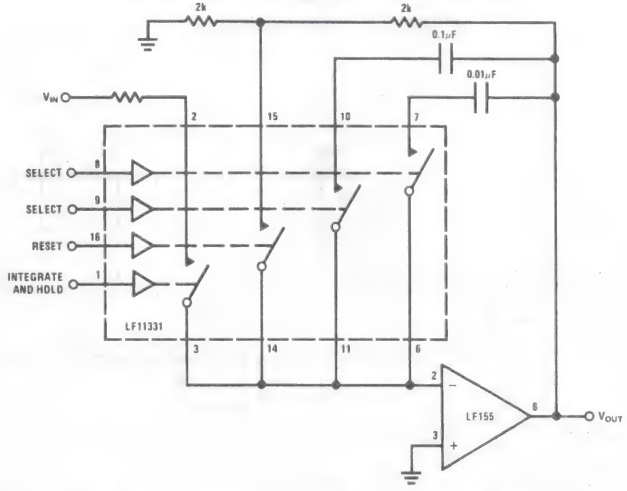
Self-Zeroing Operational Amplifier



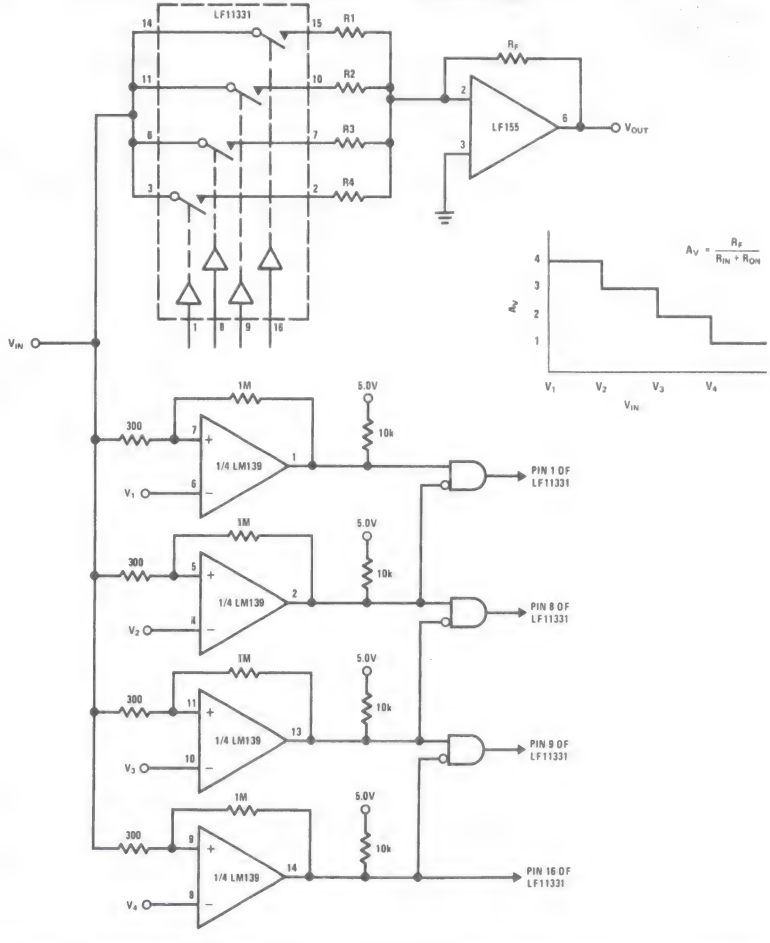
TL/H/5667–9

Typical Applications (Continued)

Programmable Integrator with Reset and Hold



Staircase Transfer Function Operational Amplifier



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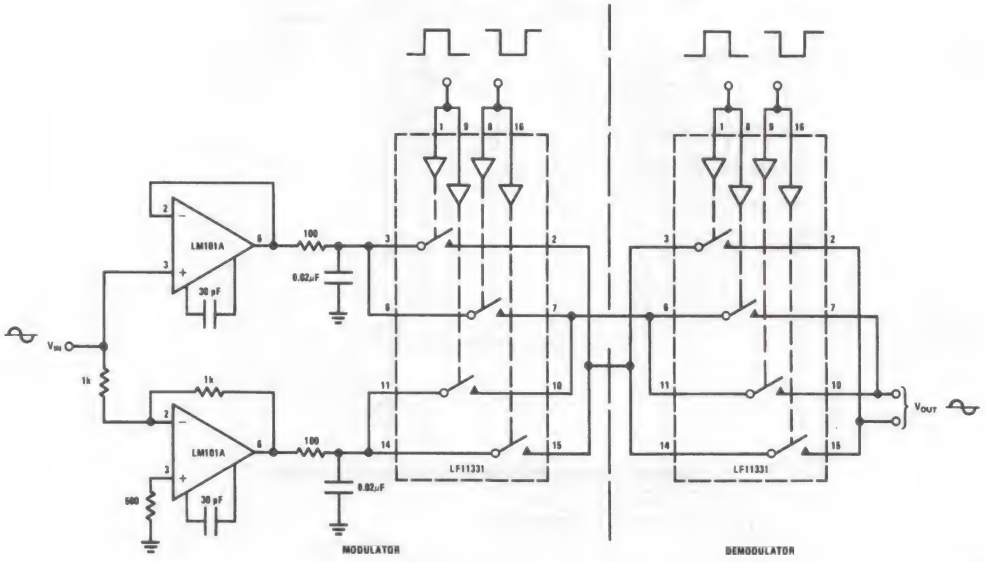
Typical Applications (Continued)

DSB Modulator-Demodulator



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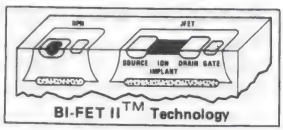
National Semiconductor



LF13508 8-Channel Analog Multiplexer
LF13509 4-Channel Differential Analog Multiplexer

General Description

The LF13508 is an 8-channel analog multiplexer which connects the output to 1 of the 8 analog inputs depending on the state of a 3-bit binary address. An enable control allows disconnecting the output, thereby providing a package select function.

This device is fabricated with National's BI-FET technology which provides ion-implanted JFETs for the analog switch on the same chip as the bipolar decode and switch drive circuitry. This technology makes possible low constant "ON" resistance with analog input voltage variations. This device does not suffer from latch-up problems or static charge blow-out problems associated with similar CMOS parts. The digital inputs are designed to operate from both TTL and CMOS levels while always providing a definite break-before-make action.

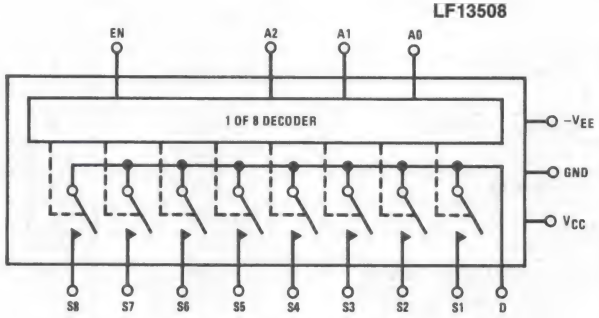
The LF13509 is a 4-channel differential analog multiplexer. A 2-bit binary address will connect a pair of independent analog inputs to one of any 4 pairs of independent analog outputs. The device has all the features of the LF13508 series and should be used whenever differential analog inputs are required.

Features

- JFET switches rather than CMOS
- No static discharge blow-out problem
- No SCR latch-up problems
- Analog signal range 11V, −15V
- Constant "ON" resistance for analog signals between −11V and 11V
- "ON" resistance 380 Ω typ
- Digital inputs compatible with TTL and CMOS
- Output enable control
- Break-before-make action: $t_{OFF} = 0.2\ \mu s$; $t_{ON} = 2\ \mu s$ typ
- Lower leakage devices available

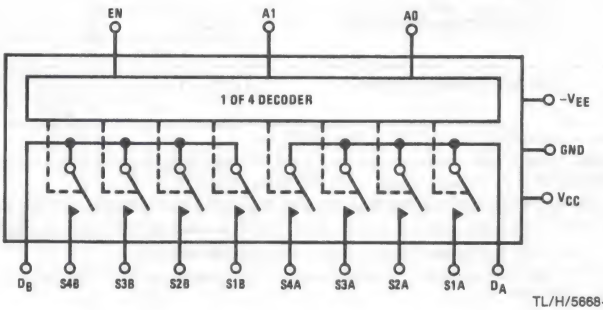
Functional Diagrams and Truth Tables

LF13508



EN	A2	A1	A0	SWITCH ON
H	L	L	L	S1
H	L	L	H	S2
H	L	H	L	S3
H	L	H	H	S4
H	H	L	L	S5
H	H	L	H	S6
H	H	H	L	S7
H	H	H	H	S8
L	X	X	X	NONE

LF13509



EN	A1	A0	SWITCH PAIR ON
L	X	X	None
H	L	L	S1
H	L	H	S2
H	H	L	S3
H	H	H	S4

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Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Positive Supply − Negative Supply ($V_{CC}-V_{EE}$)	36V
Positive Analog Input Voltage (Note 1)	V_{CC}
Negative Analog Input Voltage (Note 1)	$-V_{EE}$
Positive Digital Input Voltage	V_{CC}
Negative Digital Input Voltage	−5V
Analog Switch Current	$\|I_S\|<10$ mA

Power Dissipation (P_D at 25°C) (Notes 2 & 7)		
Molded DIP (N)	P_D	500 mW
Cavity DIP (D)	P_D	900 mW
Maximum Junction Temperature (T_{jMAX})		100°C
Operating Temperature Range		$0°C \le T_A \le +70°C$
Storage Temperature Range		−65°C to +150°C
Lead Temperature (Soldering, 10 sec.)		300°C

Electrical Characteristics (Note 3)

Symbol	Parameter	Conditions		LF13508 LF13509			Units
				Min	Typ	Max	
R_{ON}	"ON" Resistance	$V_{OUT}=0V$, $I_S=100\ \mu A$	$T_A=25°C$		380	650	Ω
					500	850	Ω
ΔR_{ON}	ΔR_{ON} with Analog Voltage Swing	$-10V \le V_{OUT} \le +10V$, $I_S=100\ \mu A$	$T_A=25°C$		0.01	1	%
R_{ON} Match	R_{ON} Match Between Switches	$V_{OUT}=0V$, $I_S=100\ \mu A$	$T_A=25°C$		20	150	Ω
$I_{S(OFF)}$	Source Current in "OFF" Condition	Switch "OFF", $V_S=11$, $V_D=-11$, (Note 4)	$T_A=25°C$			5	nA
					0.09	50	nA
$I_{D(OFF)}$	Drain Current in "OFF" Condition	Switch "OFF", $V_S=11$, $V_D=-11$, (Note 4)	$T_A=25°C$			20	nA
					0.6	500	nA
$I_{D(ON)}$	Leakage Current in "ON" Condition	Switch "ON" $V_D=11V$, (Note 4)	$T_A=25°C$			20	nA
					1	500	nA
V_{INH}	Digital "1" Input Voltage			2.0			V
V_{INL}	Digital "0" Input Voltage					0.7	V
I_{INL}	Digital "0" Input Current	$V_{IN}=0.7V$	$T_A=25°C$		1.5	30	μA
						40	μA
$I_{INL(EN)}$	Digital "0" Enable Current	$V_{EN}=0.7V$	$T_A=25°C$		1.2	30	μA
						40	μA
t_{TRAN}	Switching Time of Multiplexer	*(Figure 1)*, (Note 5)	$T_A=25°C$		1.8		μs
t_{OPEN}	Break-Before-Make	*(Figure 3)*	$T_A=25°C$		1.6		μs
$t_{ON(EN)}$	Enable Delay "ON"	*(Figure 2)*	$T_A=25°C$		1.6		μs
$t_{OFF(EN)}$	Enable Delay "OFF"	*(Figure 2)*	$T_A=25°C$		0.2		μs
$I_{SO(OFF)}$	"OFF" Isolation	(Note 6)	$T_A=25°C$		−66		dB
CT	Crosstalk	LF13509 Series, (Note 6)	$T_A=25°C$		−66		dB
$C_{S(OFF)}$	Source Capacitance ("OFF")	Switch "OFF", $V_{OUT}=0V$, $V_S=0V$	$T_A=25°C$		2.2		pF
$C_{D(OFF)}$	Drain Capacitance ("OFF")	Switch "OFF", $V_{OUT}=0V$, $V_S=0V$	$T_A=25°C$		11.4		pF
I_{CC}	Positive Supply Current	All Digital Inputs Grounded	$T_A=25°C$		7.4	12	mA
					7.9	15	mA
I_{EE}	Negative Supply Current	All Digital Inputs Grounded	$T_A=25°C$		2.7	5	mA
					2.8	6	mA

Note 1: If the analog input voltage exceeds this limit, the input current should be limited to less than 10 mA.

Note 2: The maximum power dissipation for these devices must be derated at elevated temperatures and is dictated by T_{jMAX}, θ_{jA}, and the ambient temperature, T_A. The maximum available power dissipation at any temperature is $P_D=(T_{jMAX}-T_A)/\theta_{jA}$ or the 25°C P_{DMAX}, whichever is less.

Note 3: These specifications apply for $V_S=\pm15V$ and over the absolute maximum operating temperature range ($T_L \le T_A \le T_H$) unless otherwise noted.

Note 4: Conditions applied to leakage tests insure worse case leakages. Exceeding 11V on the analog input may cause an "OFF" channel to turn "ON".

Note 5: Lots are sample tested to this parameter. The measurement conditions of *Figure 1* insure worse case transition time.

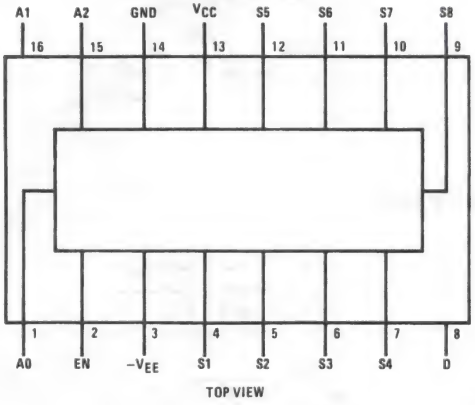
Note 6: "OFF" isolation is measured with all switches "OFF" and driving a source. Crosstalk is measured with a pair of switches "ON", driving channel A and measuring channel B. $R_L=200$, $C_L=7$ pF, $V_S=3$ Vrms, $f=500$ kHz.

Note 7: Thermal Resistance θ_{jA} (Junction to Ambient)

Molded DIP (N)	150°C/W
Cavity DIP (D)	100°C/W

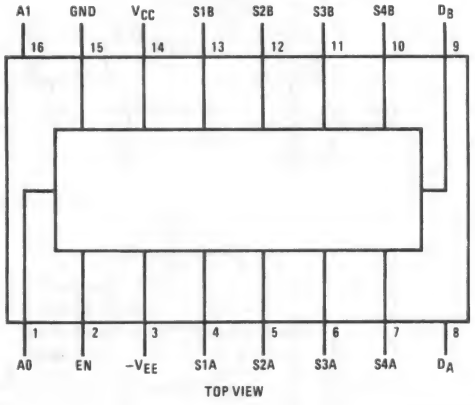
Connection Diagrams

LF13508
Dual-In-Line Package



Order Number LF13508D
See NS Package Number D16C
Order Number LF13508N
See NS Package Number N16A

LF13509
Dual-In-Line Package



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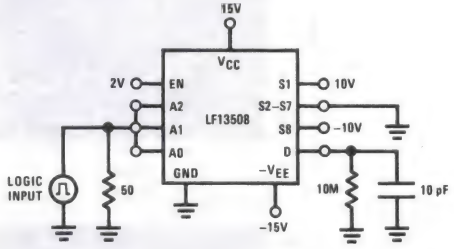
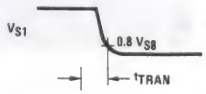
Order Number LF13509D
See NS Package Number D16C
Order Number LF13509N
See NS Package Number N16A

AC Test Circuits and Switching Time Waveforms

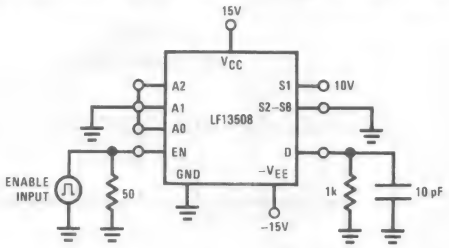




TL/H/5668–3

FIGURE 1. Transition Time

AC Test Circuit and Switching Time Waveforms (Continued)



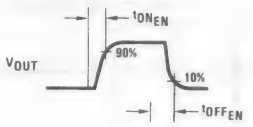
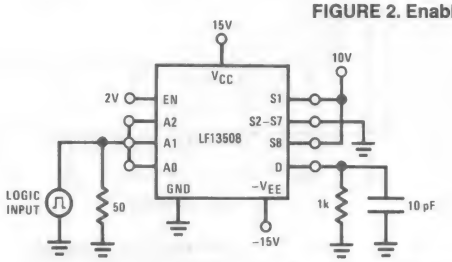
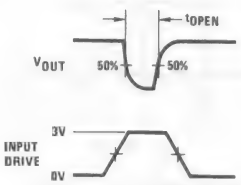


FIGURE 2. Enable Times



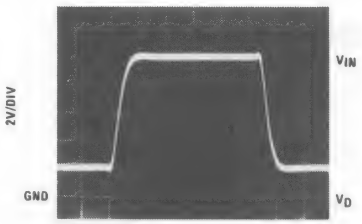


TL/H/5668–4

FIGURE 3. Break-Before-Make

Transition Times and Transients

V_A = 10V



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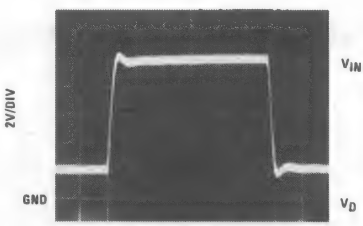
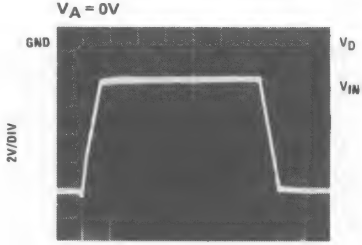
V_A = 5V



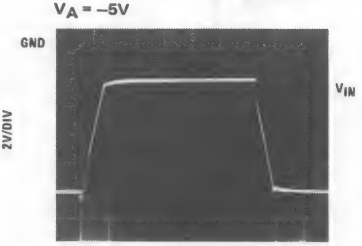
TL/H/5668–6

V_A = 0V



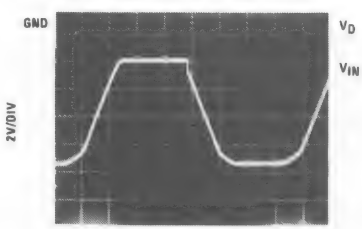
TL/H/5668–7

V_A = –5V



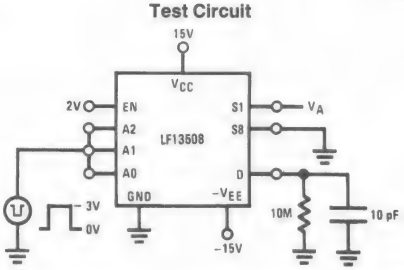
TL/H/5668–8

V_A = –10V



TL/H/5668–9

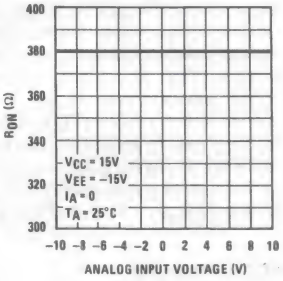
Test Circuit



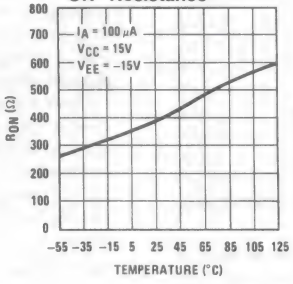
TL/H/5668–10

Typical Performance Characteristics

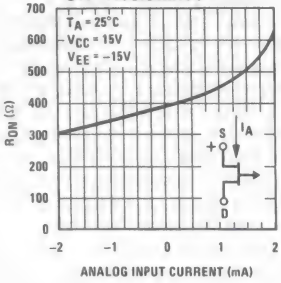
"ON" Resistance



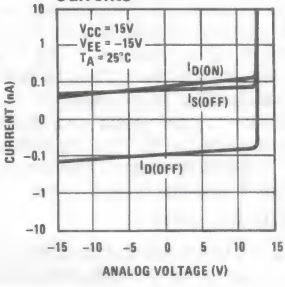
"ON" Resistance



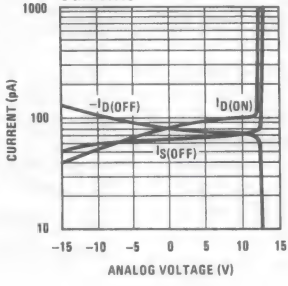
"ON" Resistance



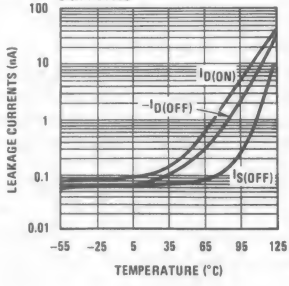
Switch Leakage Currents



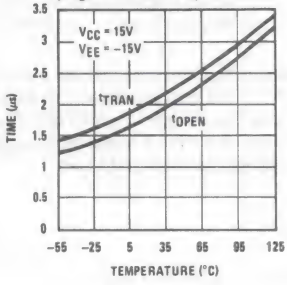
Switch Leakage Currents



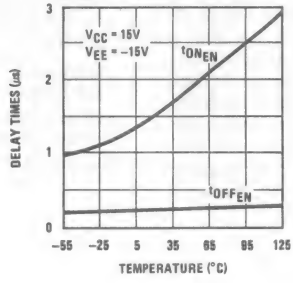
Switch Leakage Currents



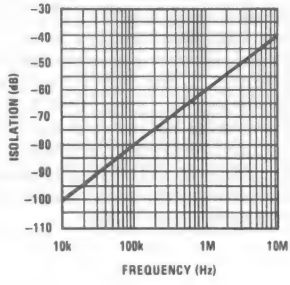
Switching Times (Figures 1 and 3)



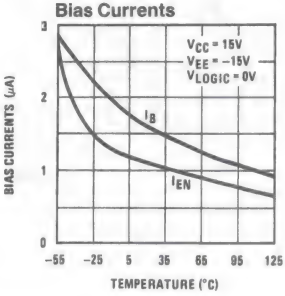
Enable Delay Times (Figure 2)



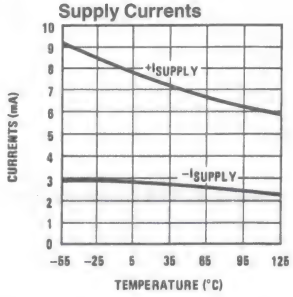
"OFF" Isolation and Crosstalk



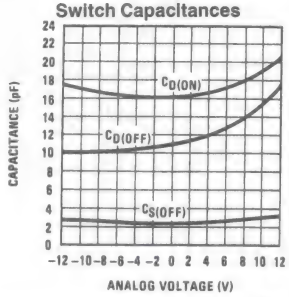
Bias Currents



Supply Currents



Switch Capacitances



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2

Application Hints

The LF11508 series is an 8-channel analog multiplexer which allows the connection of a single load to 1 of 8 different analog inputs. These multiplexers incorporate JFETs in a switch configuration which insures a constant "ON" resistance over the analog voltage range of the device. Four TTL compatible inputs are provided; a 3-bit binary decode to select a particular channel and an enable input used as a package select. The switches operate with a break-before-make action preventing the temporary connection of 2 analog inputs during switching. Because these multiplexers are fabricated with the BI-FET process rather than CMOS, they do not require special handling.

The LF11509 series is a 4-channel differential multiplexer which allows two loads to be connected to 1 of 4 different pairs of analog inputs. The LF11509 series also has all the features of the LF11508.

ANALOG VOLTAGE AND CURRENT

The "ON" resistance, R_{ON}, of the analog switches is constant over a wide input range from positive (V_{CC}) supply to negative ($-V_{EE}$) supply.

The analog input should not exceed either positive or negative supply without limiting the current to less than 10 mA; otherwise the multiplexer may get damaged. For proper operation, however, the positive analog voltage should be kept equal to or less than $V_{CC} - 4V$ as this will increase the switch leakage in both "ON" and "OFF" state and it may also cause a false turn "ON" of a normally "OFF" switch. This limit applies over the full temperature range.

The maximum allowable switch "ON" voltage (the drop across the switch in the "ON" condition) is $\pm 0.4V$ over temperature. If this number is to exceed the input current should be limited to 10 mA.

The "ON" resistance of the multiplexing switches varies slightly with analog current because they are JFETs running at 0V gate to source. The JFET characteristics shown in *Figure 4* indicates how R_{ON} tends to vary with current. A lower R_{ON} is possible when the source voltage is negative with respect to the drain voltage because the JFET becomes enhanced. Caution should be used when operating in this mode as this may forward-bias an internal transistor and cause high currents to flow in the switches. Thus, the drain voltage should never be greater than 0.4V positive with respect to the source voltage without limiting the drain current to less than 10 mA.

LEAKAGE CURRENTS

Leakage currents will remain within the specified value as long as the drain and source remain within the specified analog voltage range. As the switch terminals exceed the positive analog voltage range "ON" and "OFF" leakage currents increase. The "ON" leakage increases due to an internal clamp required by the switch structure. The "OFF" leakage increases because the gate to source reverse bias has been decreased to the point where the switch becomes active. Leakage currents vary slightly with analog voltage and will approximately double for every 10°C rise in temperature.

SWITCHING TIMES AND TRANSIENTS

These multiplexers operate with a break-before-make switch action. The turn off time is much faster than the turn on time to guarantee this feature over the full range of analog input voltage and temperature. Switching transients are introduced when a switch is turned "OFF". The amplitude of these transients may be reduced by increasing the load capacitance or decreasing the load resistance. The actual charge transfer in the transient may be reduced by operating on reduced power supplies. Examples of switching times and transients are shown in the typical characteristic curves. The enable function switching times are specified separately from switch-to-switch transition times and may be thought of as package-to-package transition times.

LOGIC INPUTS AND ENABLE INPUT

Switch selection in the LF11508 series is accomplished by using a 3-bit binary decode while the LF11509 series uses a 2-bit decode. These binary logic inputs are compatible with both TTL and CMOS logic voltage levels. The maximum positive voltage applied to these inputs may exceed V_{CC} but should not exceed $-V_{EE}+36V$. The maximum negative voltage should not be less than 4V below ground as this will cause an internal device to zener and all the switches will turn "ON".

As shown in the schematic diagram, the logic low bias current will flow until the PNP input is raised above the 3 diode reference ($\approx 2.1V$). Above this voltage the input device becomes reverse biased and the input current drops to the leakage of the reverse biased junction ($<0.1\ \mu A$).

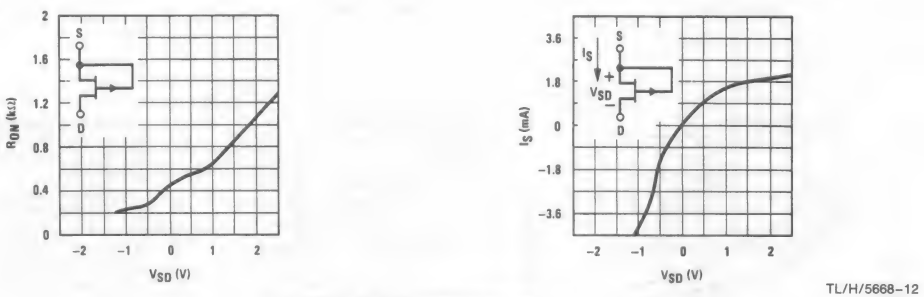

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FIGURE 4. JFET Characteristics

Typical Applications

DATA ACQUISITION SYSTEM

A SIMPLIFIED SYSTEM DISCUSSION

Analog multiplexers (MUX) are usually used for multi-channel Data Acquisition Units (DAU). *Figure 5* shows a system in which 8 different analog inputs are sampled and converted into digital words for further processing. The sample and hold circuit is optional, depending on input speed requirements and on A/D converter speed.

Parameters characterizing the system are:

System Channels: The number of multiplexer channels.

Accuracy: The conversion accuracy of each individual sample with the system operating at the throughput rate.

Speed or Throughput Rate: Number of samples/second/channel the system can handle.

For a discussion on system structure, addressing mode and processor interfacing, see application note AN-159.

A. ACCURACY CONSIDERATIONS

1. Multiplexer's Influence on System Accuracy *(Figure 6)*.

   a. The error, (E), caused by the finite "ON" resistance, R_{ON}, of the multiplexing switches is given by:

$$E(\%) = \frac{100}{1 + R_{IN}/(R_{ON} + R_S + \Delta R_{ON})}$$ where:

   R_{IN} = following stage input impedance

   ΔR_{ON} = "ON" resistance modulation which is negligible for JFET switches like the LF11508

   Example: Let R_{ON} = 450 Ω, ΔR_{ON} = 0, R_S = 0, T_A = 25°C and allowable E = 0.01% which is equivalent to 1/2 LSB in a 12-bit system:

$$R_{IN}\Big|_{min} = \frac{R_{ON}(100 - E)}{E} = 4.5\ M\Omega$$

   Note that if temperature effects are included, some gain (or full scale) drift will occur; but effects on linearity are small.

   b. Multiplexer settling time (t_s):

   $t_{s(ON)}$: is the time required for the MUX output to settle within a predetermined accuracy, as shown in Table I.

   C_S *(Figure 6):* MUX output capacitance + following stage input capacitance + any stray capacitance at this node.

TABLE I.

ERROR %	BITS	t_s(ON) TO 1/2 LSB
0.2	8	6.2t
0.05	10	7.6t
0.01	12	9t
0.0008	16	11.8t

$$t = C_S (R_{ON} + R_S) \parallel R_{IN}$$

   $t_{s(OFF)}$: is the time it takes to discharge C_S within a tolerable error. The "OFF" settling time should be taken into account for bipolar inputs where its effects will appear as a worse case of doubling of the $t_{s(ON)}$.

2. Sample and Hold Influence on System Accuracy

   The sample and hold, if used, also introduces errors into the system accuracy due to:
   - Offset voltage of sample and hold
   - Droop rate in the Hold mode
   - T_A: Aperture time or time delay between the time of a digital Hold command and the actual Hold occurance
   - Taq: Acquisition time or time it takes to acquire an analog input and settle within a predetermined error band
   - Hold step: Error created during the Sample to Hold mode caused by an undesirable charge injected into the Hold capacitor C_h.

   For more details on sample and hold errors, see the LF198/LF298/LF398 data sheet.

3. A/D Converter Influence on System Accuracy

   The "accuracy" of the A/D converter is the best possible system accuracy. In most data acquisition systems, the A/D converter is the most expensive single component, so its error will often dominate system error. Care should be taken that MUX, S/H and input source errors do not exceed system error requirements when *added* to A/D errors. For instance, if an 8-bit accuracy system is desired and an 8-bit A/D converter is used, the accuracy of the MUX and S/H should be far better than 8 bits.

For details on A/D converter specifications, see AN-156.

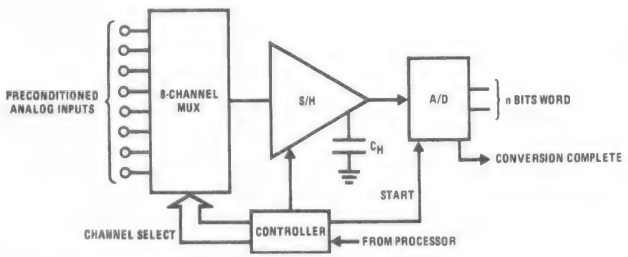

FIGURE 5. Random-Addressed, Multiplexed DAU

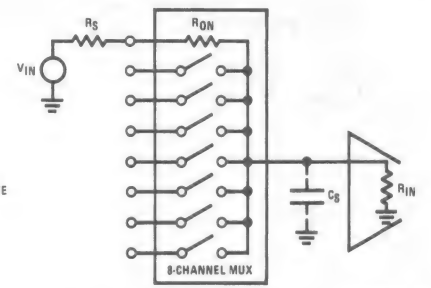

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FIGURE 6. 8-Channel MUX

Typical Applications (Continued)

B. SPEED CONSIDERATIONS

In the system of *Figure 5* with the S/H omitted, if n-bit accuracy is desired, the change of the analog input voltage should be less than ±1/2 LSB over the A/D conversion time T_C. In other words, the analog input slew rate, (rate of change of input voltage), will cause a slew-induced error and its magnitude, with respect to the total system error, will depend on the particular application.

$$\left.\frac{\Delta V_{IN}}{\Delta t}\right|_{max} < \frac{\pm 1/2\ LSB}{T_C} = \frac{V_{FS}}{2^n \times T_C}$$

where V_{FS} is the full scale voltage of the A/D. Note that slew induced errors are not affected by the MUX switch time since we can let the unit settle before starting conversion.

Example: Let $T_C = 40\ \mu s$ (MM4357), $V_{FS} = 10V$ and n = 8.

$$\left.\frac{\Delta V_{IN}}{\Delta t}\right|_{max} < \frac{1mV}{\mu s}$$

which is a very small number. A 10 Vp-p sine wave of a frequency greater than 32 Hz will have higher slew rate than this. The maximum throughput rate of the above 8-channel system would be calculated using both the A/D conversion time and the sum of MUX switch "ON" time and settling time, i.e.:

$$\text{Th. R}\Big|_{max} = \frac{1}{8(T_C + T_{MUX})} = 3k\ \text{samples/sec/channel}$$

$$T_{MUX} = T_{ON} + T_{S(ON)}$$

Also notice that Nyquist sampling criteria would allow each channel to have a signal bandwidth of 1.5 kHz max, while the slew limit dictates a maximum frequency of 32 Hz. If the input signal has a peak-to-peak voltage less than 10V, the allowable maximum input frequency can be calculated by:

$$f_{MAX} = \frac{(\text{Slew Rate})max}{\pi\ Vp\text{-}p}$$

On the other hand, if the input voltage is not band-limited a low pass filter with an attenuation of 30 dB or better at 1.5 kHz, should be connected in front of the MUX.

1. Improving System Speed with a Sample and Hold

The system speed can be improved by using the S/H shown in *Figure 5*. This allows a much greater rate of change of V_{IN}.

$$\left.\frac{\Delta V_{IN}}{\Delta t}\right|_{max} < \frac{V_{FS}}{2^n \times T_A}$$

where T_A is the aperture time of the S/H. This represents an input slew rate improvement by a factor: T_C/T_A. Here again, the slew rate error is not affected by the acquisition time of the Sample and Hold since conversion will start after the S/H has settled. *An important thing to notice is that the sample and hold errors will add to the total system error budget; therefore, the inequality of the $\Delta V_{IN}/\Delta t$ expression should become more stringent.*

Example: $T_C = 40\ \mu s$, $T_A = 0.5\ \mu s$, n = 8: $T_C/T_A = 80$

So the use of a S/H allows a speed improvement by nearly two orders of magnitude.

The maximum throughput rate can be calculated by:

$$\text{Th. R}\Big|_{max} = \frac{1}{8(T_A + Taq + T_C)}$$

Notice that T_{MUX} does not affect the $\Delta V_{IN}/\Delta t$ expression *nor the throughput rate* of the system since it may be switched and settled while the Sample and Hold is in the Hold mode. This is true, provided that: $T_{MUX} < T_A + T_C$.

C. SYSTEM EXAMPLE *(Figure 7)*

The LF398 S/H with a 1000 pF hold capacitor, has an acquisition time of 4 μs to 0.1% (1/4 LSB error for 8 bits) and an aperture time of less than 200 μs. On the other hand, after the hold command, the output will settle to ±0.05 mV in 1 μs. This, together with the acquisition time, introduces approximately a ±1/4 LSB error. Allowing another 1/4 LSB error for hold step and gain non-linearity, the maximum slew error ($\Delta V_{IN}/\Delta t$) should not exceed 1/4 LSB or:

$$\frac{\Delta V_{IN}}{\Delta t} \le \frac{1}{4} \times \frac{1}{256} \times \frac{1}{T_A} \approx 5mV/\mu s$$

(which is the maximum slew rate of a 5 V peak sine wave. Also notice that, due to the above input slew restrictions, the analog delay caused by the finite BW of the S/H and the digital delay caused by the response time of the controller will be negligible. The maximum throughput rate of the system is:

$$\text{Th. R}\Big|_{max} = \frac{1}{8(5 + 40)10^{-6}} = 2800\ \text{samples/sec/ch.}$$

If the system speed requirements are relaxed, but the A/D converter is still too slow, then an inexpensive S/H can be built by using just a capacitor and a low cost FET input op amp as shown in *Figure 8.*

Typical Applications (Continued)

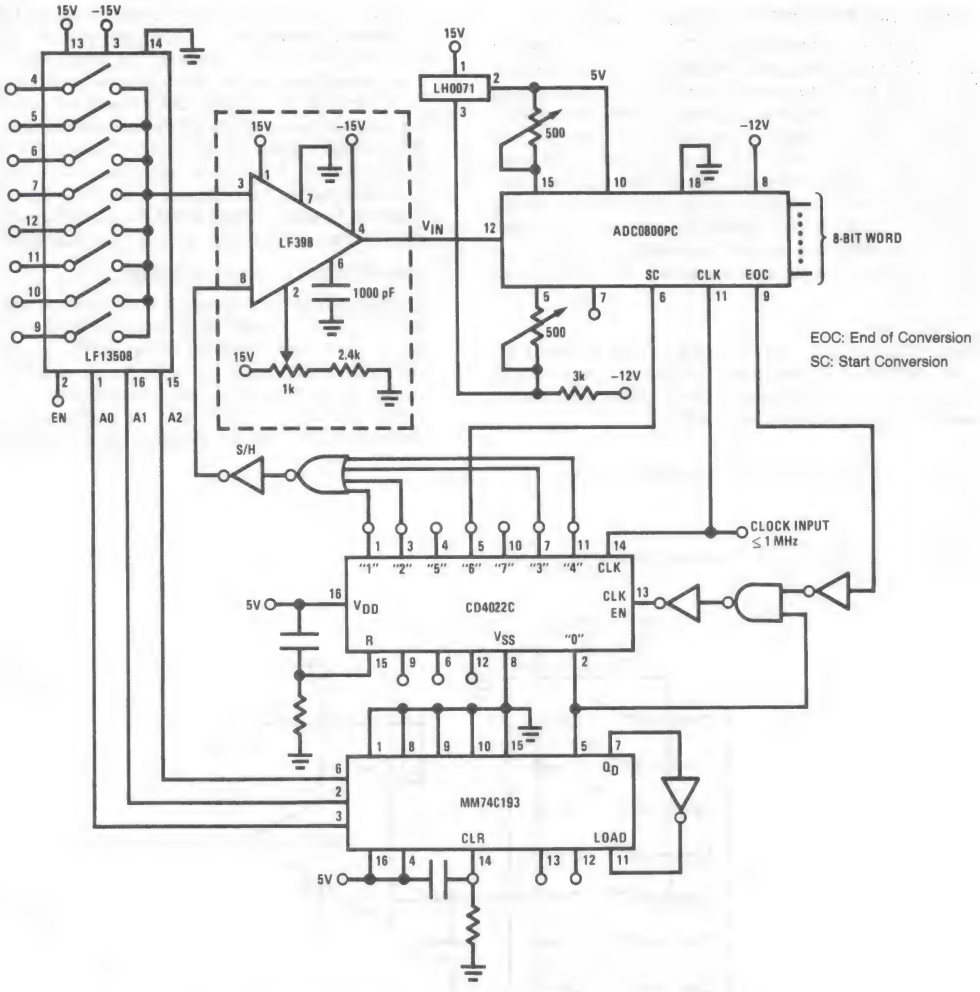

FIGURE 7a. Sequentially Multiplexed DAU with Sample and Hold

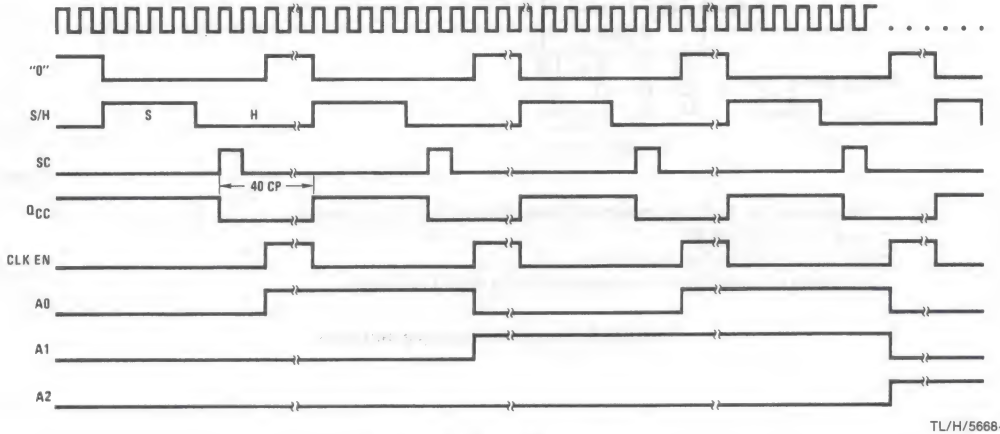

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FIGURE 7b. Timing Diagram

Typical Applications (Continued)

D. DOUBLING THE SYSTEM CHANNEL CAPABILITY

This is done in two different ways. First, we can use second level multiplexing with speed benefits, as shown in *Figure 9.* A fast 2-channel multiplexer, made by the dual analog switch AM182, accepts the outputs of each 8-channel MUX, LF13508, and then feeds them sequentially into an 8-bit successive approximation A/D converter. With this technique, the throughput rate of the system can again be made independent of the LF13508 speed. Looking at the timing diagram, when the A/D converter converts the analog value of an upper multiplexer channel, we switch channels in the lower multiplexer for the next conversion. This can be done provided that:

$$T_{MUX} \leq T_C + 1\ CP$$

The LF356 connected as unity gain buffers are used because of the low input impedance of the A/D; they are connected between multiplexers for speed optimization. With a maximum clock frequency of 4.5 MHz:

$$\text{Th. R} = \frac{10^6}{16 \times 2} = 31.25\text{k samples/sec/channel}$$

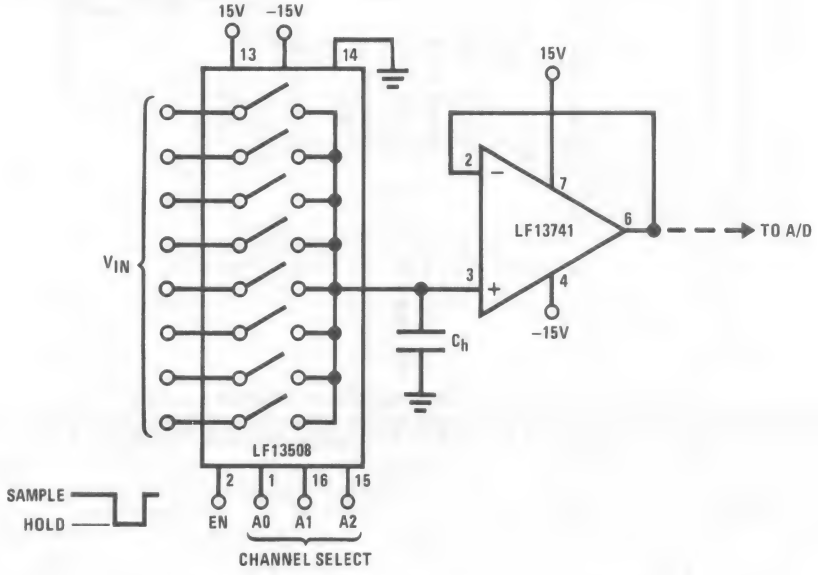
and

$$\left.\frac{\Delta V_{IN}}{\Delta t}\right|_{max} < \frac{10}{256} \times \frac{1}{2\mu s} = 19.5\ mV/\mu s \text{ for } 10V_{FS}$$

An alternate way to increase the system channel is shown in *Figure 10,* where the enable pins are used to disable one MUX while the other is sampling. With this method, many 8-channel multiplexers can be connected, but the parasitic capacitance at the common output node will keep increasing and will eventually degrade the settling time, $t_{s(ON)}$. Also, the MUX speed will now affect the system throughput. If, for instance, this method was used instead of second level multiplexing, the system of *Figure 9* will lose half of its speed. If, however, speed is not the prime system requirement, the approach of *Figure 10* is more cost effective.

E. DIFFERENTIAL INPUT SYSTEMS

Systems operating in industrial environments may require an instrumentation amplifier to separate the desired analog signal from any common-mode signal present. The LF11509 was designed to provide 4 pairs of differential input signals to the input of an instrumentation amplifier for further process. A 4-channel preconditioning circuit is shown in *Figure 11* and a complete system is shown in *Figure 12.*



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- The acquisition time, T_A, of the Sample and Hold depends upon: R_{ON}, I_{DSS} of switches, Z_{OUT} of switches
- $I_{DSS} \cong 1.5$ mA, $Z_{OUT} = 40\ k\Omega$
- $V_{IN} = 10V$, $C_h = 1000$ pF, $T_A = 20\ \mu s$ to 0.1%
- Error created by charge injection during Hold mode: $\Delta V_E \cong 10\ pF\ (14.5V - V_{IN})/C_h$

FIGURE 8. Inexpensive Sample and Hold

Typical Applications (Continued)

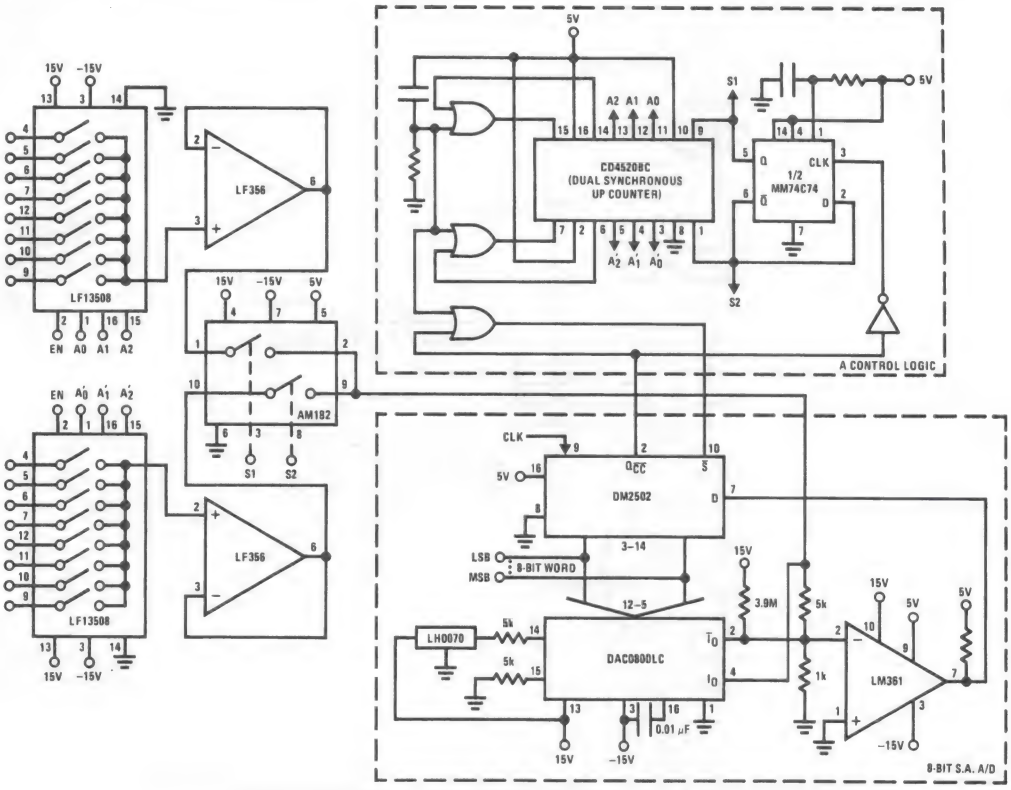

FIGURE 9a. A Fast 16-Channel DAU with Second Level Multiplexing

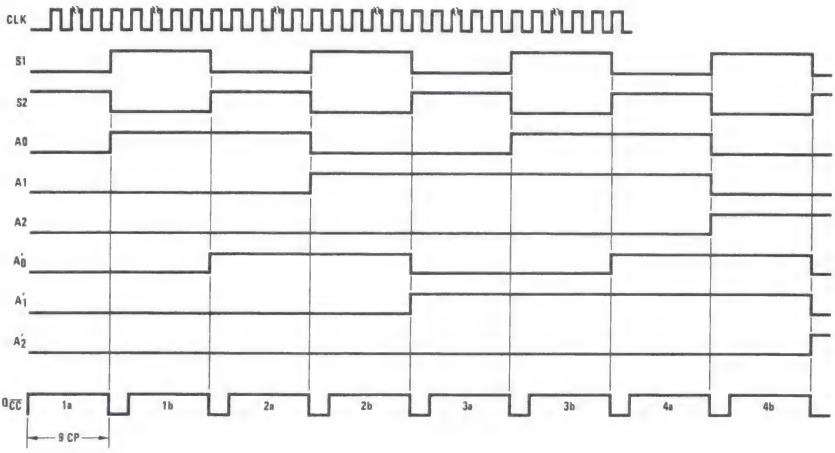

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FIGURE 9b. Timing Diagram

Typical Applications (Continued)

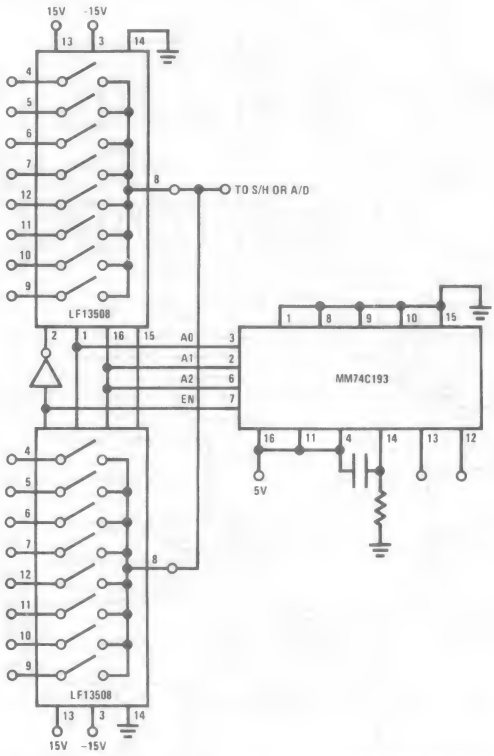

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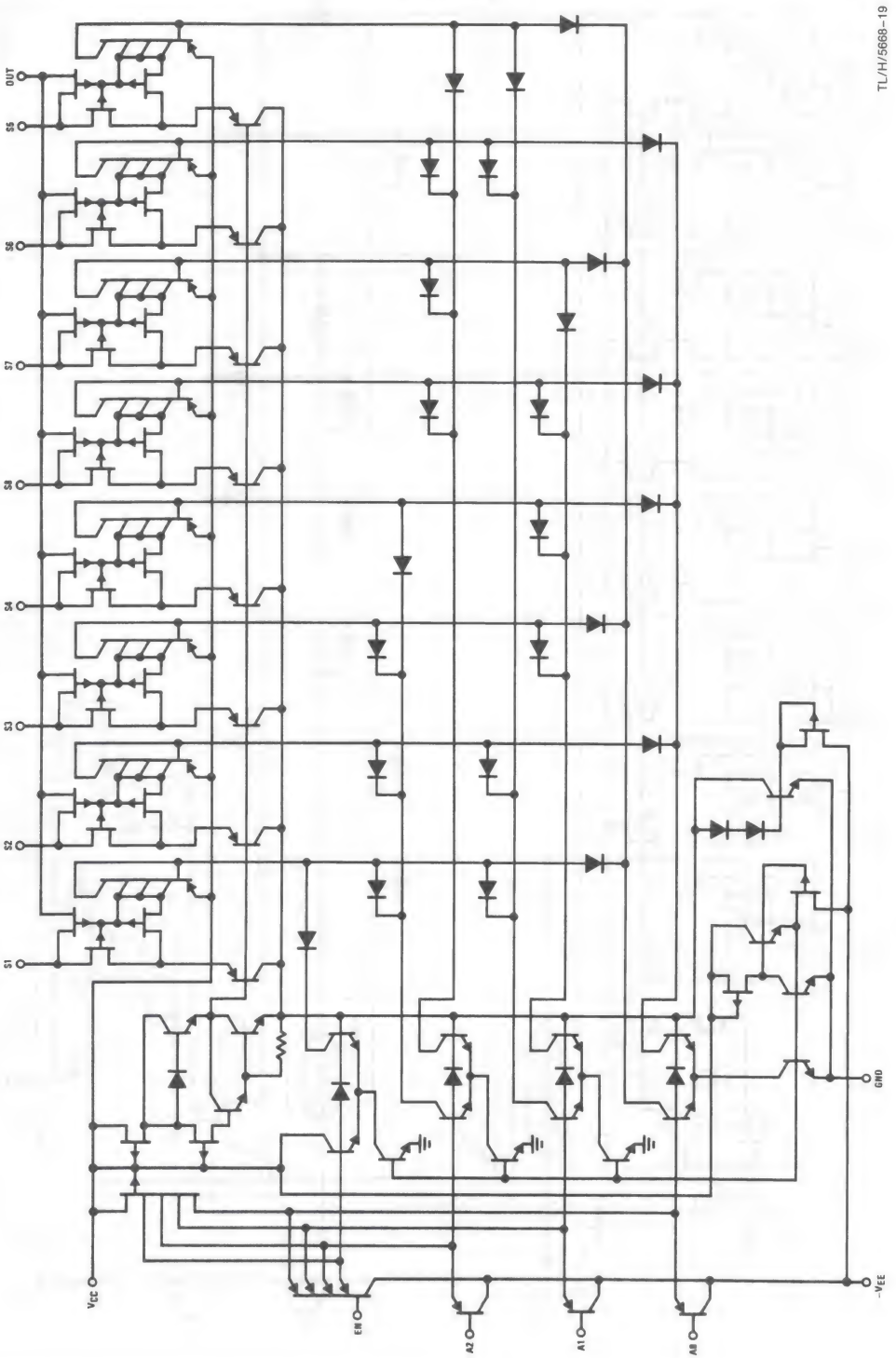
FIGURE 10. A 16-Channel Multiplexer with Sequential Multiplexing

Schematic Diagrams

LF13508



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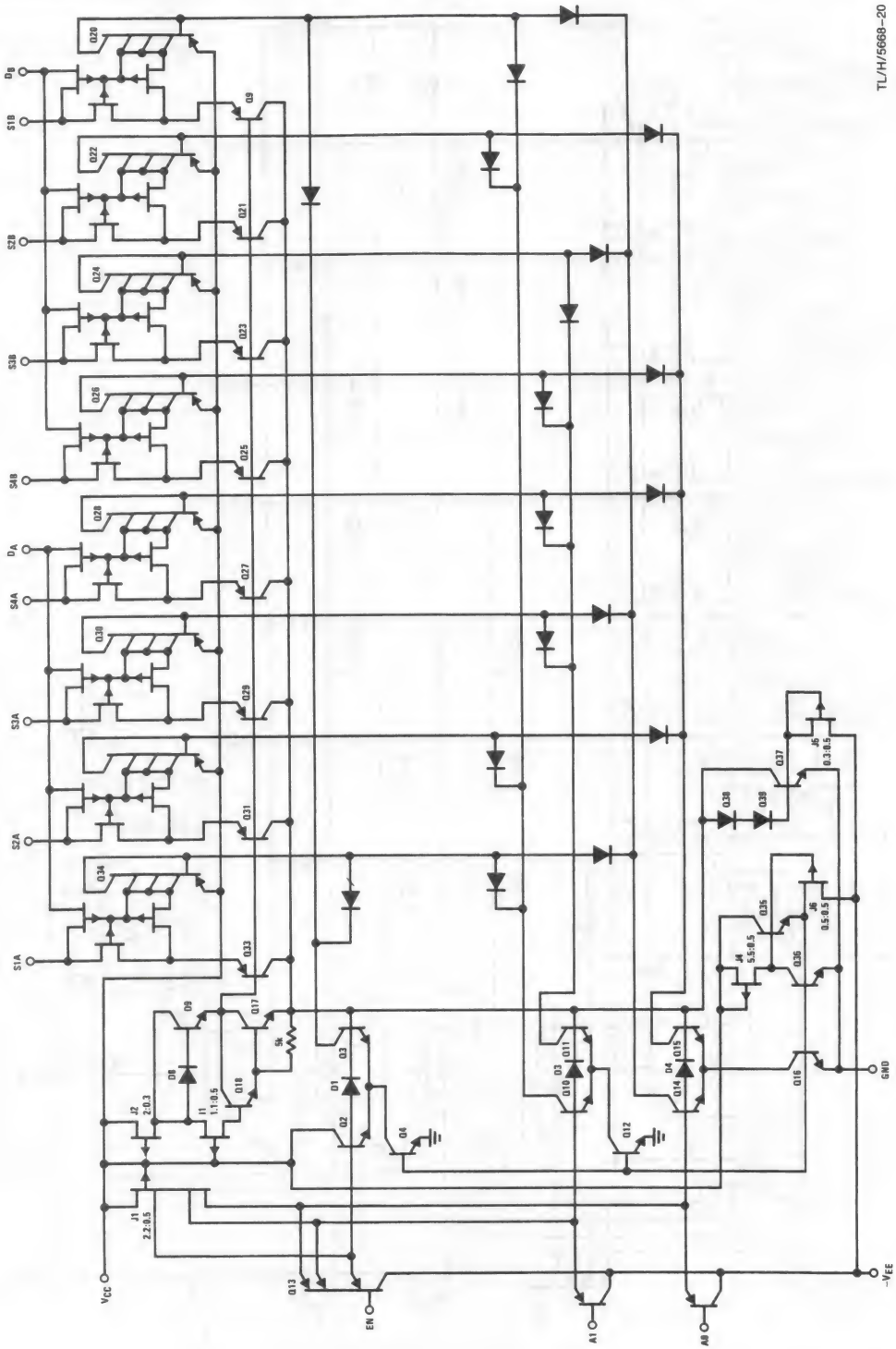
Schematic Diagrams (Continued)

LF13509



National Semiconductor

MM54HC4016/MM74HC4016 Quad Analog Switch

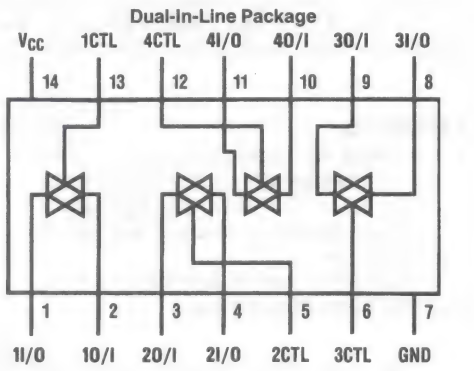
General Description

These devices are digitally controlled analog switches implemented in advanced silicon-gate CMOS technology. These switches have low "on" resistance and low "off" leakages. They are bidirectional switches, thus any analog input may be used as an output and vice-versa. The '4016 devices allow control of up to 12V (peak) analog signals with digital control signals of the same range. Each switch has its own control input which disables each switch when low. All analog inputs and outputs and digital inputs are protected from electrostatic damage by diodes to V_{CC} and ground.

Features

- Typical switch enable time: 15 ns
- Wide analog input voltage range: 0–12V
- Low "on" resistance: 50Ω typ.
- Low quiescent current: 80 μA maximum (74HC)
- Matched switch characteristics
- Individual switch controls

Connection Diagram



Top View

Order Number MM54HC4016* or MM74HC4016*

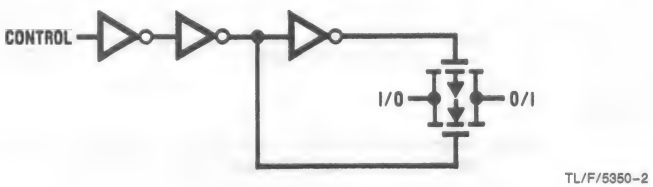
*Please look into Section 8, Appendix D for availability of various package types.

Truth Table

Input	Switch
CTL	I/O-O/I
L	"OFF"
H	"ON"

See the CMOS Logic Databook for complete specifications

Schematic Diagram



National Semiconductor

MM54HC4051/MM74HC4051 8-Channel Analog Multiplexer
MM54HC4052/MM74HC4052 Dual 4-Channel Analog Multiplexer
MM54HC4053/MM74HC4053 Triple 2-Channel Analog Multiplexer

General Description

These multiplexers are digitally controlled analog switches implemented in advanced silicon-gate CMOS technology. These switches have low "on" resistance and low "off" leakages. They are bidirectional switches, thus any analog input may be used as an output and vice-versa. Also these switches contain linearization circuitry which lowers the on resistance and increases switch linearity. These devices allow control of up to ±6V (peak) analog signals with digital control signals of 0 to 6V. Three supply pins are provided for V_{CC}, ground, and V_{EE}. This enables the connection of 0–5V logic signals when $V_{CC}=5V$ and an analog input range of ±5V when $V_{EE}=5V$. All three devices also have an inhibit control which when high will disable all switches to their off state. All analog inputs and outputs and digital inputs are protected from electrostatic damage by diodes to V_{CC} and ground.

MM54HC4051/MM74HC4051: This device connects together the outputs of 8 switches, thus achieving an 8 channel Multiplexer. The binary code placed on the A, B, and C select lines determines which one of the eight switches is "on", and connects one of the eight inputs to the common output.

MM54HC4052/MM74HC4052: This device connects together the outputs of 4 switches in two sets, thus achieving a pair of 4-channel multiplexers. The binary code placed on the A, and B select lines determine which switch in each 4 channel section is "on", connecting one of the four inputs in each section to its common output. This enables the implementation of a 4-channel differential multiplexer.

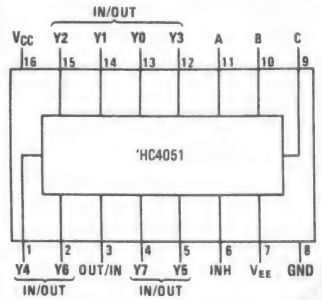
MM54HC4053/MM74HC4053: This device contains 6 switches whose outputs are connected together in pairs, thus implementing a triple 2 channel multiplexer, or the equivalent of 3 single-pole-double throw configurations. Each of the A, B, or C select lines independently controls one pair of switches, selecting one of the two switches to be "on".

Features

- Wide analog input voltage range: ±6V
- Low "on" resistance: 50 typ. ($V_{CC}-V_{EE}=4.5V$)
  30 typ. ($V_{CC}-V_{EE}=9V$)
- Logic level translation to enable 5V logic with ±5V analog signals
- Low quiescent current: 80 μA maximum (74HC)
- Matched Switch characteristic

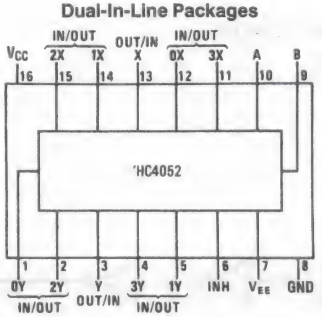
Connection Diagrams

Dual-In-Line Packages



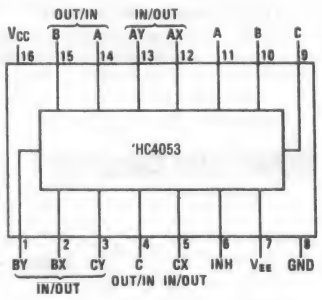
TL/F/5353–1

Top View



TL/F/5353–2

Top View



TL/F/5353–3

Top View

Order Number MM54HC4051*, MM74HC4051*, MM54HC4052*, MM74HC4052*, MM54HC4053* or MM74HC4053*

*Please look into Section 8, Appendix D for availability of various package types.

See the CMOS Logic Databook for Complete Specifications

National Semiconductor

MM54HC4066/MM74HC4066 Quad Analog Switch

General Description

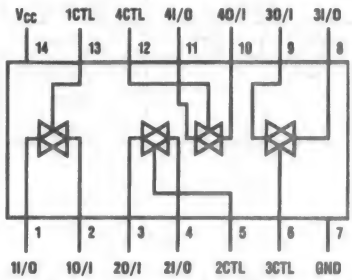
These devices are digitally controlled analog switches utilizing advanced silicon-gate CMOS technology. These switches have low "on" resistance and low "off" leakages. They are bidirectional switches, thus any analog input may be used as an output and visa-versa. Also the '4066 switches contain linearization circuitry which lowers the "on" resistance and increases switch linearity. The '4066 devices allow control of up to 12V (peak) analog signals with digital control signals of the same range. Each switch has its own control input which disables each switch when low. All analog inputs and outputs and digital inputs are protected from electrostatic damage by diodes to V_{CC} and ground.

Features

- Typical switch enable time: 15 ns
- Wide analog input voltage range: 0–12V
- Low "on" resistance: 30 typ. ('4066)
- Low quiescent current: 80 μA maximum (74HC)
- Matched switch characteristics
- Individual switch controls

Connection Diagram

Dual-In-Line Package



TL/F/5355–1

Top View

Order Number MM54HC4066* or MM74HC4066*

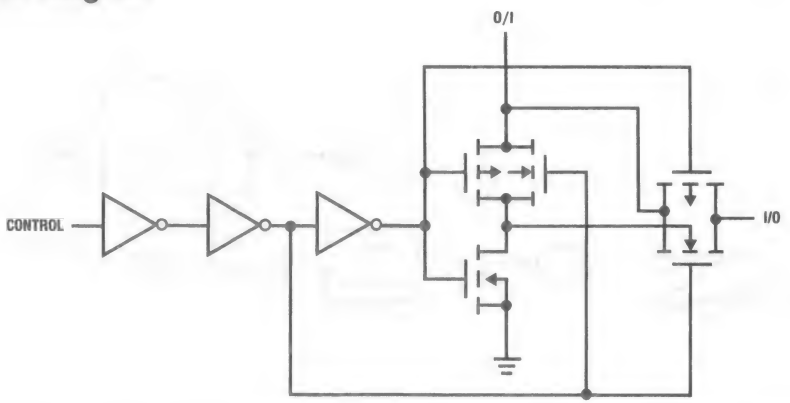
*Please look into Section 8, Appendix D for availability of various package types.

Truth Table

Input	Switch
CTL	I/O–O/I
L	"OFF"
H	"ON"

See the CMOS Logic Databook for Complete Specifications

Schematic Diagram



TL/F/5355–2

National Semiconductor

MM54HC4316/MM74HC4316
Quad Analog Switch with Level Translator

General Description

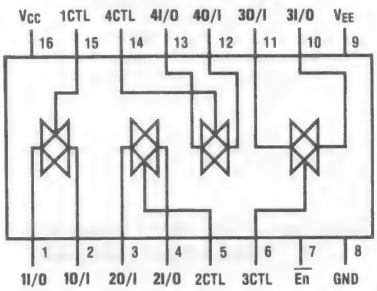
These devices are digitally controlled analog switches implemented in advanced silicon-gate CMOS technology. These switches have low "on" resistance and low "off" leakages. They are bidirectional switches, thus any analog input may be used as an output and vice-versa. Three supply pins are provided on the '4316 to implement a level translator which enables this circuit to operate with 0–6V logic levels and up to ±6V analog switch levels. The '4316 also has a common enable input in addition to each switch's control which when low will disable all switches to their off state. All analog inputs and outputs and digital inputs are protected from electrostatic damage by diodes to V_{CC} and ground.

Features

- Typical switch enable time: 20 ns
- Wide analog input voltage range: ±6V
- Low "on" resistance: 50 typ. ($V_{CC}-V_{EE}=4.5V$)
  30 typ. ($V_{CC}-V_{EE}=9V$)
- Low quiescent current: 80 μA maximum (74HC)
- Matched switch characteristics
- Individual switch controls plus a common enable

Connection and Logic Diagrams

Dual-In-Line Package



TL/F/5369–1

Top View

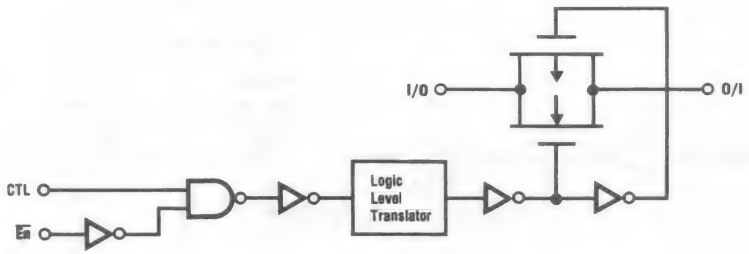
Order Number MM54HC4316* or MM74HC4316*

*Please look into Section 8, Appendix D for availability of various package types.

Truth Table

Inputs		Switch
$\overline{En}$	CTL	I/O–O/I
H	X	"OFF"
L	L	"OFF"
L	H	"ON"

See the CMOS Logic Databook for Complete Specifications



TL/F/5369–2

Section 3

Analog-to-Digital Converters

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Definition Of Terms A/D Converters

Conversion Time: The time required for a complete measurement by an analog-to-digital converter.

DC Common-Mode Error: This specification applies to ADCs with differential inputs. It is the change in the output code that occurs when the analog voltages on the two inputs are changed by an equal amount. It is expressed in LSBs.

Differential Nonlinearity: Ideally, any two adjacent digital codes correspond to measured analog voltages that are exactly one LSB apart. Differential non-linearity is a measure of the worst case deviation from the ideal 1 LSB step. For example, a DAC with a 1.5 LSB output change for a 1 LSB digital code change exhibits ½ LSB differential non-linearity. Differential non-linearity may be expressed in fractional bits or as a percentage of full scale. A differential non-linearity greater than 1 LSB will lead to a non-monotonic transfer function in a DAC and missing codes in an ADC.

Gain Error (Full Scale Error): For an ADC, the difference (usually expressed in LSBs) between the input voltage that should ideally produce a full scale output code and the actual input voltage that produces that code. For DACs, it is the difference between the output voltage (or current) with full scale input code and the ideal voltage (or current) that should exist with a full scale input code.

Gain Temperature Coefficient (Full Scale Temperature Coefficient): Change in gain error divided by change in temperature. Usually expressed in parts per million per degree Celsius (ppm/°C).

Integral Nonlinearity (Linearity Error): Worst case deviation from the line between the endpoints (zero and full scale). Can be expressed as a percentage of full scale or in fractions of an LSB.

LSB (Least-Significant Bit): In a binary coded system this is the bit that carries the smallest value or weight. Its value is the full scale voltage (or current) divided by 2^n, where n is the resolution of the converter.

Missing Codes: When an incremental increase or decrease in input voltage causes the converter to increment or decrement its numeric output by more than one LSB the converter is said to exhibit "missing codes". If there are missing codes, there is a numeric value on the output on the converter which cannot be reached by any input voltage value.

Monotonicity: A monotonic function has a slope whose sign does not change. A monotonic DAC has an output that changes in the same direction (or remains constant) for each increase in the input code. The converse is true for decreasing codes.

MSB (Most Significant Bit): In a binary coded system this is the bit that has the largest value or weight. Its value is one half of full scale.

Multiplying DAC: In a sense, every DAC is a multiplying DAC since the output voltage (or current) is equal to the reference voltage times a constant determined by the digital input code divided by 2^n (n is the number of bits of resolution). In a two quadrant multiplying DAC the reference voltage or the digital input code can change the output voltage polarity. If both the reference voltage and the digital code change the output voltage polarity four quadrant multiplication exists.

Offset Error (Zero Error): In a DAC, this is the output voltage that exists when the input digital code is set to give an ideal output of zero volts. In the case of an ADC, this is the difference between the ideal input voltage (½ LSB) and the actual input voltage that is needed to make the transition from zero to 1 LSB. All the digital codes in the transfer curve are offset by the same value. Many converters allow nulling of offset with an external potentiometer. Offset error is usually expressed in LSBs.

Power Supply Rejection (Power Supply Sensitivity): The sensitivity of a converter to changes in the dc power supply voltages.

Quantizing Error: The error inherent in all A/D conversions. Since even an "ideal" converter has finite resolution, any analog voltage that falls between two adjacent output codes will result in an output code that is inaccurate by up to ½ LSB.

Ratiometric Operation: Many A/D applications require a stable and accurate reference voltage against which the input voltage is compared. This approach results in an **absolute conversion**. Some applications, however, use transducers or other signal sources whose output voltages are proportional to some external reference. In these **ratiometric** applications, the reference for the signal source should be connected to the reference input of the converter. Thus, any variations in the source reference voltage will also change the converter reference voltage and produce an accurate conversion.

Resolution: The smallest analog increment corresponding to a 1 LSB converter code change. For converters, resolution is normally expressed in bits, where the number of analog levels is equal to 2^n. As an example, a 12-bit converter divides the analog signal into $2^{12} = 4096$ discrete voltage (or current) levels.

Settling Time: The time from a change in input code until a DAC's output signal remains within ± ½ LSB (or some other specified tolerance) of the final value.

National Semiconductor

A/D Converter Selection Guide

Part No.	Resolution (Bits)	Accuracy (Max)	Conversion Time	Input Voltage Range	Output Logic Levels	Supplies (V)	Temperature Range*			Package	Comments
							M	I	C		
A/D CONVERTER											
†ADC0800	8	±2 LSB	50 μs	±5V	TTL, TRI-STATE	+5, −12	•		•	18-Pin DIP	
†ADC0801	8	±¼ LSB	110 μs	5V	TTL, TRI-STATE	+5	•	•		20-Pin DIP	Differential Input
†ADC0802	8	±½ LSB	110 μs	5V	TTL, TRI-STATE	+5	•	•	•	20-Pin DIP 20-Pin SO 20-Pin PCC	Differential Input
†ADC0803	8	±½ LSB	110 μs	5V	TTL, TRI-STATE	+5	•	•	•	20-Pin DIP 20-Pin SO 20-Pin PCC	Differential Input
†ADC0804	8	±1 LSB	110 μs	5V	TTL, TRI-STATE	+5		•	•	20-Pin DIP 20-Pin SO 20-Pin PCC	Differential Input
†ADC0805	8	±1 LSB	110 μs	5V	TTL, TRI-STATE	+5		•		20-Pin DIP	Ratiometric Operation
†ADC0808	8	±½ LSB	100 μs	5V	TTL, TRI-STATE	+5	•	•		28-Pin DIP 28-Pin PCC	8-Channel MUX
†ADC0809	8	±1 LSB	100 μs	5V	TTL, TRI-STATE	+5		•		28-Pin DIP 28-Pin PCC	8-Channel MUX
†ADC0811B	8	±½ LSB	32 μs	5V	TTL	+5		•	•	20-Pin DIP 20-Pin PCC	11-Channel Serial I/O
†ADC0811C	8	±1 LSB	32 μs	5V	TTL	+5		•	•	20-Pin DIP 20-Pin PCC	11-Channel Serial I/O
†ADC0816	8	±½ LSB	100 μs	5V	TTL, TRI-STATE	+5	•	•		40-Pin DIP	16-Channel MUX
†ADC0817	8	±1 LSB	100 μs	5V	TTL, TRI-STATE	+5		•		40-Pin DIP	16-Channel MUX
†ADC0819B	8	±½ LSB	16 μs	5V	TTL	+5		•	•	28-Pin DIP 28-Pin PCC	19-Channel Serial I/O
†ADC0819C	8	±1 LSB	16 μs	5V	TTL	+5		•	•	28-Pin DIP 28-Pin PCC	19-Channel Serial I/O
†ADC0820B	8	±½ LSB	1.2 μs	5V	TTL, TRI-STATE	+5	•	•	•	20-Pin DIP 20-Pin SO 20-Pin PCC	Built-In Track and Hold Function
†ADC0820C	8	±1 LSB	1.2 μs	5V	TTL, TRI-STATE	+5	•	•	•	20-Pin DIP 20-Pin SO 20-Pin PCC	Built-In Track and Hold Function

A/D Converter Selection Guide (Continued)

Part No.	Resolution (Bits)	Accuracy (Max)	Conversion Time	Input Voltage Range	Output Logic Levels	Supplies (V)	Temperature Range*			Package	Comments
							M	I	C		
A/D CONVERTER (Continued)											
†ADC0829B	8	±½ LSB	100 μs	5V	TTL, TRI-STATE	+5		•		28-Pin DIP	Additional Digital Input Capability
†ADC0829C	8	±1 LSB	100 μs	5V	TTL, TRI-STATE	+5		•		28-Pin DIP	Additional Digital Input Capability
†ADC0831B	8	±½ LSB	32 μs	5V	TTL	+5		•	•	8-Pin DIP	Serial I/O
†ADC0831C	8	±1 LSB	32 μs	5V	TTL	+5		•	•	8-Pin DIP	Serial I/O
†ADC0832B	8	±½ LSB	32 μs	5V	TTL	+5		•	•	8-Pin DIP	2-Channel Serial I/O
†ADC0832C	8	±1 LSB	32 μs	5V	TTL	+5		•	•	8-Pin DIP	2-Channel Serial I/O
†ADC0833B	8	±½ LSB	32 μs	5V	TTL	+5		•	•	14-Pin DIP	4-Channel Serial I/O
†ADC0833C	8	±1 LSB	32 μs	5V	TTL	+5		•	•	14-Pin DIP	4-Channel Serial I/O
†ADC0834B	8	±½ LSB	32 μs	5V	TTL	+5		•	•	14-Pin DIP	4-Channel Serial I/O
†ADC0834C	8	±1 LSB	32 μs	5V	TTL	+5		•	•	14-Pin DIP	4-Channel Serial I/O
†ADC0838B	8	±½ LSB	32 μs	5V	TTL	+5		•	•	20-Pin DIP 20-Pin PCC	8-Channel Serial I/O
†ADC0838C	8	±1 LSB	32 μs	5V	TTL	+5		•	•	20-Pin DIP 20-Pin PCC	8-Channel Serial I/O
†ADC0841B	8	±½ LSB	40 μs	5V	TTL, TRI-STATE	+5		•	•	20-Pin DIP 20-Pin PCC	Differential Input, Internal Clock
†ADC0841C	8	±1 LSB	40 μs	5V	TTL, TRI-STATE	+5		•	•	20-Pin DIP 20-Pin PCC	Differential Input, Internal Clock
†ADC0844B	8	±½ LSB	40 μs	5V	TTL, TRI-STATE	+5		•	•	20-Pin DIP	4-Channel MUX, Internal Clock
†ADC0844C	8	±1 LSB	40 μs	5V	TTL, TRI-STATE	+5		•	•	20-Pin DIP	4-Channel MUX, Internal Clock
†ADC0848B	8	±½ LSB	40 μs	5V	TTL, TRI-STATE	+5		•	•	24-Pin DIP 28-Pin PCC	8-Channel MUX, Internal Clock
†ADC0848C	8	±1 LSB	40 μs	5V	TTL, TRI-STATE	+5		•	•	24-Pin DIP 28-Pin PCC	8-Channel MUX, Internal Clock
ADC1001C	10	±1 LSB	200 μs	5V	TTL, TRI-STATE	+5		•	•	20-Pin DIP	8-Bit Bus Compatible, Differential Input
ADC1005B	10	±½ LSB	50 μs	5V	TTL, TRI-STATE	+5	•	•	•	20-Pin DIP 20-Pin PCC	8-Bit Bus Compatible, Differential Input

A/D Converter Selection Guide (Continued)

Part No.	Resolution (Bits)	Accuracy (Max)	Conversion Time	Input Voltage Range	Output Logic Levels	Supplies (V)	Temperature Range*			Package	Comments
							M	I	C		
A/D CONVERTER (Continued)											
ADC1005C	10	±1 LSB	50 μs	5V	TTL, TRI-STATE	+5	•	•	•	20-Pin DIP 20-Pin PCC	8-Bit Bus Compatible, Differential Input
ADC1021C	10	±1 LSB	200 μs	5V	TTL, TRI-STATE	+5		•	•	24-Pin DIP	Differential Input
ADC1025B	10	±½ LSB	50 μs	5V	TTL, TRI-STATE	+5	•	•	•	24-Pin DIP 28-Pin PCC	Differential Input
ADC1025C	10	±1 LSB	50 μs	5V	TTL, TRI-STATE	+5	•	•	•	24-Pin DIP 28-Pin PCC	Differential Input
ADC1205B	12+sign	±½ LSB	100 μs	±5V	TTL, TRI-STATE	+5, ±5		•	•	24-Pin DIP	8-Bit Bus Compatible, Differential Input
ADC1205C	12+sign	±1 LSB	100 μs	±5V	TTL, TRI-STATE	+5, ±5		•	•	24-Pin DIP	8-Bit Bus Compatible, Differential Input
ADC1210	12	±¾ LSB	200 μs	10.2V	CMOS	+5 to ±15	•	•		24-Pin DIP	Bipolar or Unipolar Input
ADC1211	12	±2 LSB	200 μs	10.2V	CMOS	+5 to ±5	•	•		24-Pin DIP	Bipolar or Unipolar Input
ADC1225B	12+sign	±½ LSB	100 μs	±5V	TTL, TRI-STATE	+5, ±5		•	•	28-Pin DIP	Differential Input
ADC1225C	12+sign	±1 LSB	100 μs	±5V	TTL, TRI-STATE	+5, ±5		•	•	28-Pin DIP	Differential Input
ADC3511	3½-Digit	0.05%	200 ms	2V	TTL, TRI-STATE	+5			•	24-Pin DIP	Integrating μP Compatible
ADC3711	3¾-Digit	0.05%	400 ms	2V	TTL, TRI-STATE	+5			•	24-Pin DIP	Integrating μP Compatible
LM131	V-F	0.01%	N/A	V_{CC} − 2V	Open Collector	+5 to +40	•	•	•	8-Pin DIP or TO-99 Can	Voltage-to-Frequency Converter 100 kHz Max
DIGITAL VOLTMETER											
ADD3501	3½-Digit	0.05%	200 ms	2V	7-Segment LED Drive	+5			•	28-Pin DIP	3½-Digit LED DVM
ADD3701	3½-Digit	0.05%	400 ms	2V	7-Segment LED Drive	+5			•	28-Pin DIP	3¾-Digit LED DVM

*Temperature ranges: "M" is −55°C to +125°C ambient; "I" is −40°C to +85°C or −25°C to +85°C; "C" is 0°C to +70°C.

†Accuracy specified is absolute accuracy and includes total unadjusted error.

National Semiconductor

ADC0800 8-Bit A/D Converter

General Description

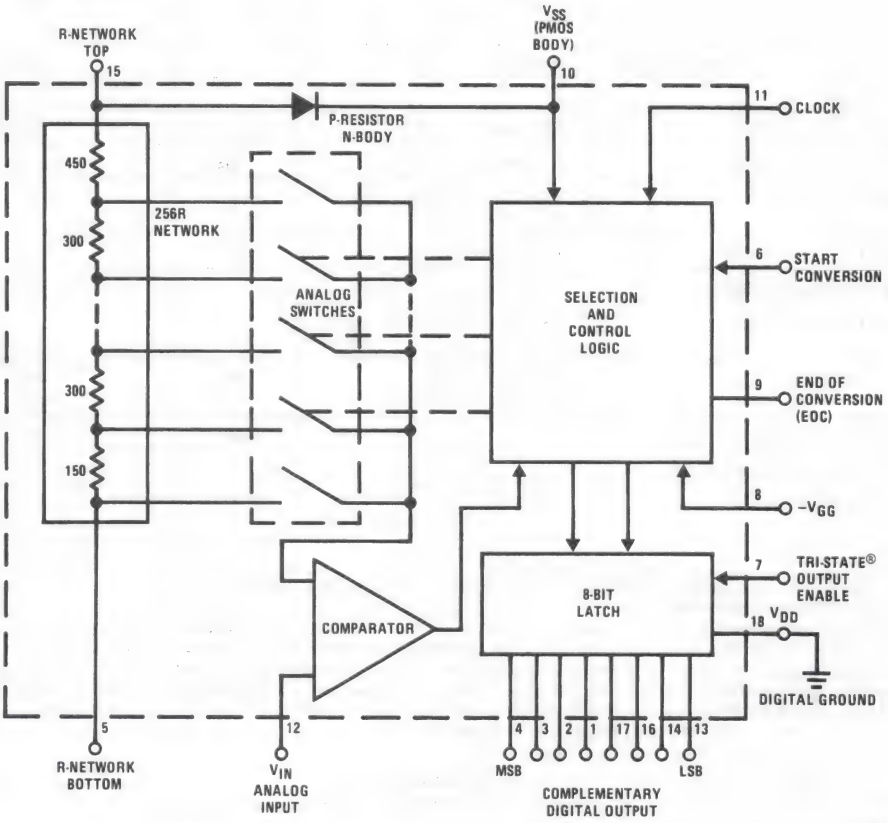
The ADC0800 is an 8-bit monolithic A/D converter using P-channel ion-implanted MOS technology. It contains a high input impedance comparator, 256 series resistors and analog switches, control logic and output latches. Conversion is performed using a successive approximation technique where the unknown analog voltage is compared to the resistor tie points using analog switches. When the appropriate tie point voltage matches the unknown voltage, conversion is complete and the digital outputs contain an 8-bit complementary binary word corresponding to the unknown. The binary output is TRI-STATE® to permit bussing on common data lines.

The ADC0800PD is specified over −55°C to +125°C and the ADC0800PCD is specified over 0°C to 70°C.

Features

■ Low cost	
■ ±5V, 10V input ranges	
■ No missing codes	
■ Ratiometric conversion	
■ TRI-STATE outputs	
■ Fast	$T_C = 50\ \mu s$
■ Contains output latches	
■ TTL compatible	
■ Supply voltages	5 V_{DC} and −12 V_{DC}
■ Resolution	8 bits
■ Linearity	±1 LSB
■ Conversion speed	40 clock periods
■ Clock range	50 to 800 kHz

Block Diagram



TL/H/5670–1

(00000000 = +full-scale)

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{DD})	$V_{SS}-22V$
Supply Voltage (V_{GG})	$V_{SS}-22V$
Voltage at Any Input	$V_{SS}+0.3V$ to $V_{SS}-22V$
Input Current at Any Pin (Note 2)	5 mA
Package Input Current (Note 2)	20 mA
Power Dissipation (Note 3)	875 mW
ESD Susceptibility (Note 4)	500V
Storage Temperature	150°C
Lead Temperature (Soldering, 10 sec.)	300°C

Operating Ratings (Note 1)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0800PD	$-55°C \le T_A \le +125°C$
ADC0800PCD	$0°C \le T_A \le +70°C$

Electrical Characteristics

These specifications apply for $V_{SS}=5.0\ V_{DC}$, $V_{GG}=-12.0\ V_{DC}$, $V_{DD}=0\ V_{DC}$, a reference voltage of 10.000 V_{DC} across the on-chip R-network ($V_{R\text{-}NETWORK\ TOP}=5.000\ V_{DC}$ and $V_{R\text{-}NETWORK\ BOTTOM}=-5.000\ V_{DC}$), and a clock frequency of 800 kHz. For all tests, a 475Ω resistor is used from pin 5 to $V_{R\text{-}NETWORK\ BOTTOM}=-5\ V_{DC}$. Unless otherwise noted, these specifications apply over an ambient temperature range of −55°C to +125°C for the ADC0800PD and 0°C to +70°C for the ADC0800PCD.

Parameter	Conditions	Min	Typ	Max	Units
Non-Linearity	$T_A=25°C$, (Note 8) Over Temperature, (Note 8)			±1 ±2	LSB LSB
Differential Non-Linearity				±½	LSB
Zero Error				±2	LSB
Zero Error Temperature Coefficient	(Note 9)			0.01	%/°C
Full-Scale Error				±2	LSB
Full-Scale Error Temperature Coefficient	(Note 9)			0.01	%/°C
Input Leakage				1	μA
Logical "1" Input Voltage	All Inputs	$V_{SS}-1.0$		V_{SS}	V
Logical "0" Input Voltage	All Inputs	V_{GG}		$V_{SS}-4.2$	V
Logical Input Leakage	$T_A=25°C$, All Inputs, $V_{IL}=V_{SS}-10V$			1	μA
Logical "1" Output Voltage	All Outputs, $I_{OH}=100\ \mu A$	2.4			V
Logical "0" Output Voltage	All Outputs, $I_{OL}=1.6$ mA			0.4	V
Disabled Output Leakage	$T_A=25°C$, All Outputs, $V_{OL}=V_{SS}$@10V			2	μA
Clock Frequency	$0°C \le T_A \le +70°C$ $-55°C \le T_A \le +125°C$	50 100		800 500	kHz kHz
Clock Pulse Duty Cycle		40		60	%
TRI-STATE Enable/Disable Time				1	μs
Start Conversion Pulse	(Note 10)	1		3½	Clock Periods
Power Supply Current	$T_A=25°C$			20	mA

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 5 mA or less. The 20 mA package input current limits the number of pins that can exceed the power supply boundaries with a 5 mA current limit to four.

Note 3: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 125°C$, and the typical junction-to-ambient thermal resistance of the ADC0800PD and ADC0800PCD when board mounted is 66°C/W.

Note 4: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 5: Typicals are at 25°C and represent most likely parametric norm.

Note 6: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

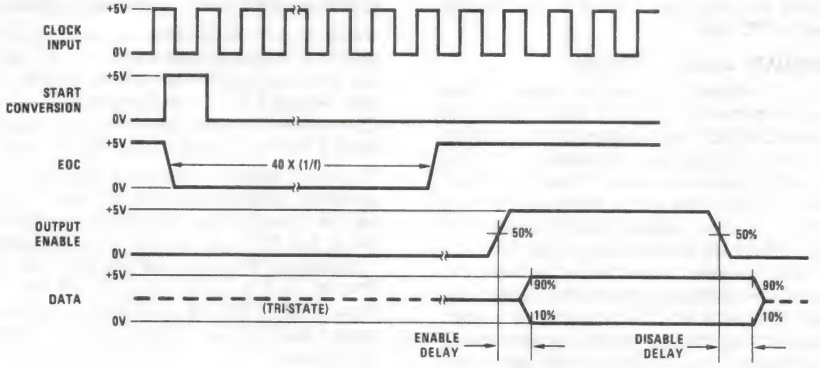
Note 7: Design limits are guaranteed but not 100% tested. These limits are not used to calculate outgoing quality levels.

Note 8: Non-linearity specifications are based on best straight line.

Note 9: Guaranteed by design only.

Note 10: Start conversion pulse duration greater than 3½ clock periods will cause conversion errors.

Timing Diagram



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Data is complementary binary (full scale is all "0's" output).

Application Hints

OPERATION

The ADC0800 contains a network with 256-300Ω resistors in series. Analog switch taps are made at the junction of each resistor and at each end of the network. In operation, a reference (10.00V) is applied across this network of 256 resistors. An analog input (V_{IN}) is first compared to the center point of the ladder via the appropriate switch. If V_{IN} is larger than $V_{REF}/2$, the internal logic changes the switch points and now compares V_{IN} and $\frac{3}{4}$ V_{REF}. This process, known as successive approximation, continues until the best match of V_{IN} and V_{REF}/N is made. N now defines a specific tap on the resistor network. When the conversion is complete, the logic loads a binary word corresponding to this tap into the output latch and an end of conversion (EOC) logic level appears. The output latches hold this data valid until a new conversion is completed and new data is loaded into the latches. The data transfer occurs in about 200 ns so that valid data is present virtually all the time in the latches. The data outputs are activated when the Output Enable is high, and in TRI-STATE when Output Enable is low. The Enable Delay time is approximately 200 ns. Each conversion requires 40 clock periods. The device may be operated in the free running mode by connecting the Start Conversion line to the End of Conversion line. However, to ensure start-up under all possible conditions, an external Start Conversion pulse is required during power up conditions.

REFERENCE

The reference applied across the 256 resistor network determines the analog input range. $V_{REF} = 10.00V$ with the top of the R-network connected to 5V and the bottom connected to −5V gives a ±5V range. The reference can be level shifted between V_{SS} and V_{GG}. However, the voltage, applied to the top of the R-network (pin 15), must not exceed V_{SS}, to prevent forward biasing the on-chip parasitic silicon diodes that exist between the P-diffused resistors (pin 15) and the N-type body (pin 10, V_{SS}). Use of a standard logic power supply for V_{SS} can cause problems, both due to initial voltage tolerance and changes over temperature. A solution is to power the V_{SS} line (15 mA max drain) from the output of the op amp that is used to bias the top of the R-network (pin 15). The analog input voltage and the voltage that is applied to the bottom of the R-network (pin 5) must be at least 7V above the $-V_{GG}$ supply voltage to ensure adequate voltage drive to the analog switches.

Other reference voltages may be used (such as 10.24V). If a 5V reference is used, the analog range will be 5V and accuracy will be reduced by a factor of 2. Thus, for maximum accuracy, it is desirable to operate with at least a 10V reference. For TTL logic levels, this requires 5V and −5V for the R-network. CMOS can operate at the 10 V_{DC} V_{SS} level and a single 10 V_{DC} reference can be used. All digital voltage levels for both inputs and outputs will be from ground to V_{SS}.

ANALOG INPUT AND SOURCE RESISTANCE CONSIDERATIONS

The lead to the analog input (pin 12) should be kept as short as possible. Both noise and digital clock coupling to this input can cause conversion errors. To minimize any input errors, the following source resistance considerations should be noted:

For $R_S \leq 5k$	No analog input bypass capacitor required, although a 0.1 μF input bypass capacitor will prevent pickup due to unavoidable series lead inductance.
For $5k < R_S \leq 20k$	A 0.1 μF capacitor from the input (pin 12) to ground should be used.
For $R_S > 20k$	Input buffering is necessary.

If the overall converter system requires lowpass filtering of the analog input signal, use a 20 kΩ or less series resistor for a passive RC section or add an op amp RC active lowpass filter (with its inherent low output resistance) to ensure accurate conversions.

CLOCK COUPLING

The clock lead should be kept away from the analog input line to reduce coupling.

LOGIC INPUTS

The logical "1" input voltage swing for the Clock, Start Conversion and Output Enable should be ($V_{SS} - 1.0V$).

Application Hints (Continued)

CMOS will satisfy this requirement but a pull-up resistor should be used for TTL logic inputs.

RE-START AND DATA VALID AFTER EOC

The EOC line (pin 9) will be in the low state for a maximum of 40 clock periods to indicate "busy". A START pulse that occurs while the A/D is BUSY will reset the SAR and start a new conversion with the EOC signal remaining in the low state until the end of this new conversion. When the conversion is complete, the EOC line will go to the high voltage state. An additional 4 clock periods must be allowed to elapse after EOC goes high, before a new conversion cycle is requested. Start Conversion pulses that occur during this last 4 clock period interval may be ignored (see *Figure 1* and *2* for high speed operation). This is a problem only for high conversion rates and keeping the number of conversions per second less than $f_{CLOCK}/44$ automatically guarantees proper operation. For example, for an 800 kHz clock, approximately 18,000 conversions per second are allowed. The transfer of the new digital data to the output is initiated when EOC goes to the high voltage state.

POWER SUPPLIES

Standard supplies are $V_{SS} = +5V$, $V_{GG} = -12V$ and $V_{DD} = 0V$. Device accuracy is dependent on stability of the reference voltage and has slight sensitivity to V_{SS}—V_{GG}. V_{DD} has no effect on accuracy. Noise spikes on the V_{SS} and V_{GG} supplies can cause improper conversion; therefore, filtering each supply with a 4.7 μF tantalum capacitor is recommended.

CONTINUOUS CONVERSIONS AND LOGIC CONTROL

Simply tying the EOC output to the Start Conversion input will allow continuous conversions, but an oscillation on this line will exist during the first 4 clock periods after EOC goes high. Adding a D flip-flop between EOC (D input) to Start Conversion (Q output) will prevent the oscillation and will allow a stop/continuous control via the "clear" input.

To prevent missing a start pulse that may occur after EOC goes high and prior to the required 4 clock period time interval, the circuit of *Figure 1* can be used. The RS latch can be set at any time and the 4-stage shift register delays the application of the start pulse to the A/D by 4 clock periods. The RS latch is reset 1 clock period after the A/D EOC signal goes to the low voltage state. This circuit also provides a Start Conversion pulse to the A/D which is 1 clock period wide.

A second control logic application circuit is shown in *Figure 2*. This allows an asynchronous start pulse of arbitrary length less than T_C, to continuously convert for a fixed high level and provides a single clock period start pulse to the A/D. The binary counter is loaded with a count of 11 when the start pulse to the A/D appears. Counting is inhibited until the EOC signal from the A/D goes high. A carry pulse is then generated 4 clock periods after EOC goes high and is used to reset the input RS latch. This carry pulse can be used to indicate that the conversion is complete, the data has transferred to the output buffers and the system is ready for a new conversion cycle.

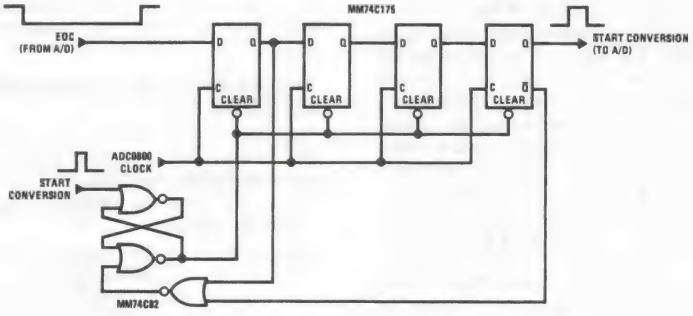

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FIGURE 1. Delaying an Asynchronous Start Pulse

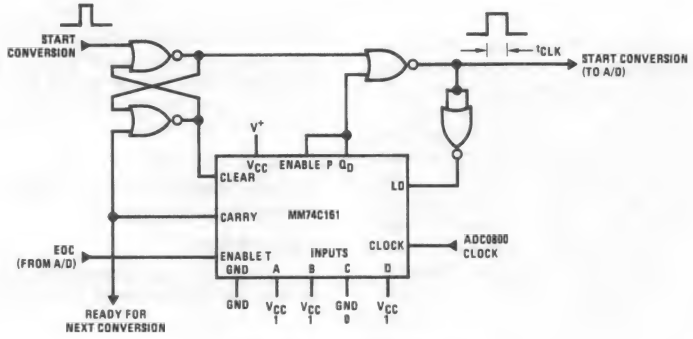

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FIGURE 2. A/D Control Logic

Application Hints (Continued)

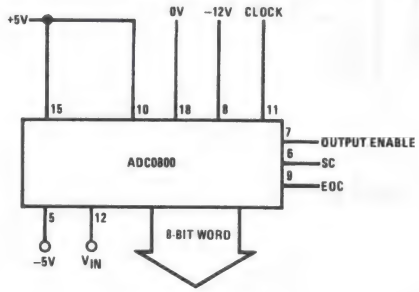
ZERO AND FULL-SCALE ADJUSTMENT

Zero Adjustment: This is the offset voltage required at the bottom of the R-network (pin 5) to make the 11111111 to 11111110 transition when the input voltage is ½ LSB (20 mV for a 10.24V scale). In most cases, this can be accomplished by having a 1 kΩ pot on pin 5. A resistor of 475Ω can be used as a non-adjustable best approximation from pin 5 to ground.

Full-Scale Adjustment: This is the offset voltage required at the top of the R-network (pin 15) to make the 00000001 to 00000000 transition when the input voltage is 1 ½ LSB from full-scale (60 mV less than full-scale for a 10.24V scale). This voltage is guaranteed to be within ±2 LSB for the ADC0800 without adjustment. In most cases, adjustment can be accomplished by having a 1 kΩ pot on pin 15.

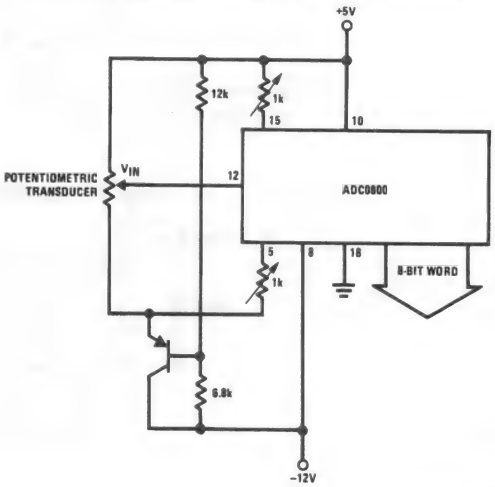
Typical Applications

General Connection



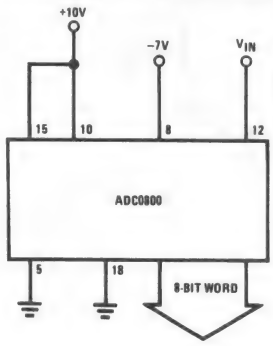
TL/H/5670–11

Ratiometric Input Signal with Tracking Reference



TL/H/5670–4

Hi-Voltage CMOS Output Levels

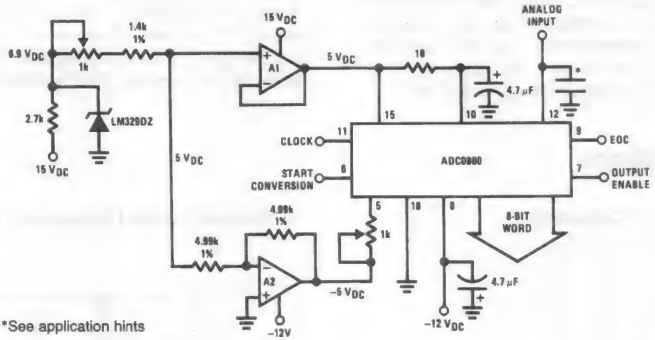


0V to 10V V_{IN} range
0V to 10V output levels

TL/H/5670–12

Typical Applications (Continued)

$V_{REF} = 10\ V_{DC}$ With TTL Logic Levels

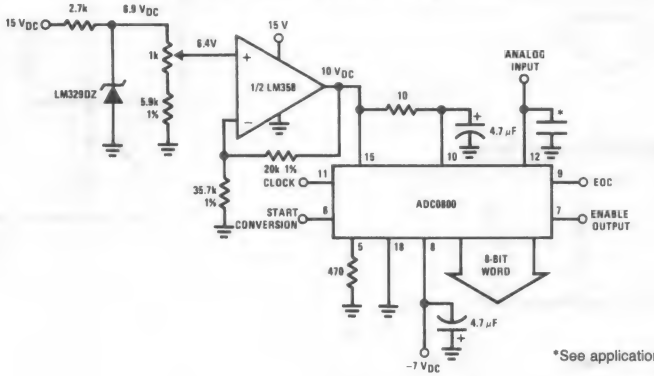


*See application hints

A1 and A2 = LM358N dual op amp

TL/H/5670–13

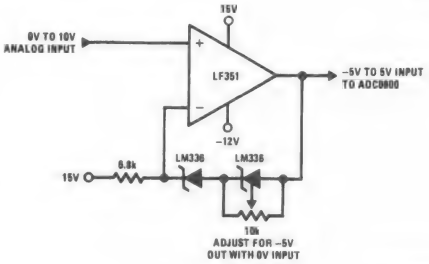
$V_{REF} = 10\ V_{DC}$ With 10V CMOS Logic Levels



*See application hints

TL/H/5670–14

Input Level Shifting



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- Permits TTL compatible outputs with 0V to 10V input range (0V to −10V input range achieved by reversing polarity of zener diodes and returning the 6.8k resistor to V−).

Typical Applications (Continued)

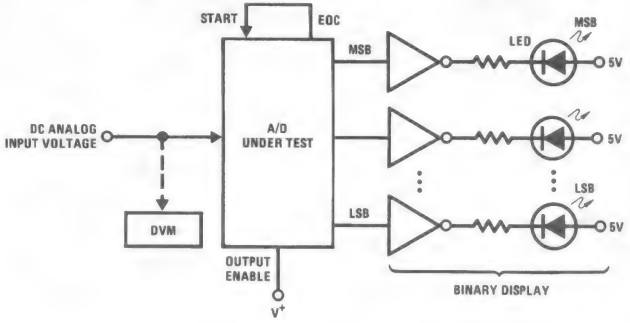
TESTING THE A/D CONVERTER

There are many degrees of complexity associated with testing an A/D converter. One of the simplest tests is to apply a known analog input voltage to the converter and use LEDs to display the resulting digital output code as shown in *Figure 3*. Note that the LED drivers invert the digital output of the A/D converter to provide a binary display. A lab DVM can be used if a precision voltage source is not available. After adjusting the zero and full-scale, any number of points can be checked, as desired.

For ease of testing, a 10.24 V_{DC} reference is recommended for the A/D converter. This provides an LSB of 40 mV (10.240/256). To adjust the zero of the A/D, an analog input voltage of ½ LSB or 20 mV should be applied and the zero adjust potentiometer should be set to provide a flicker on the LSB LED readout with all the other display LEDs OFF.

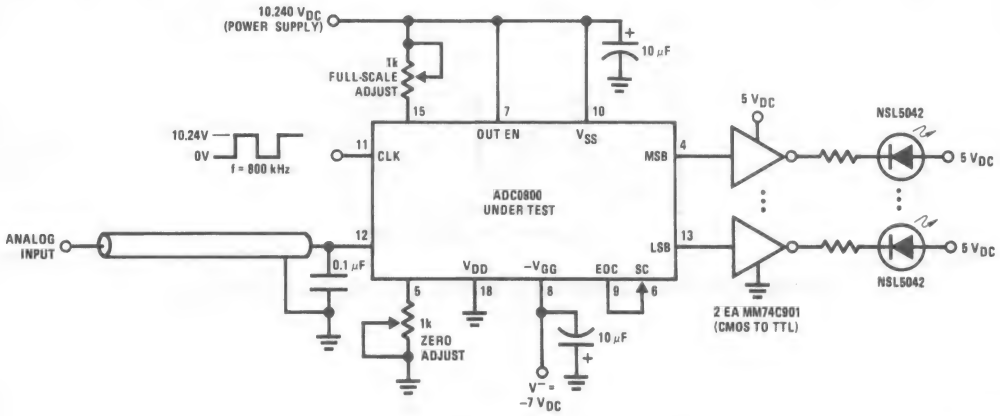
To adjust the full-scale adjust potentiometer, an analog input that is 1½ LSB less than the reference (10.240–0.060 or 10.180 V_{DC}) should be applied to the analog input and the full-scale adjusted for a flicker on the LSB LED, but this time with all the other LEDs ON.

A complete circuit for a simple A/D tester is shown in *Figure 4*. Note that the clock input voltage swing and the digital output voltage swings are from 0V to 10.24V. The MM74C901 provides a voltage translation to 5V operation and also the logic inversion so the readout LEDs are in binary.



TL/H/5670–15

FIGURE 3. Basic A/D Tester



TL/H/5670–7

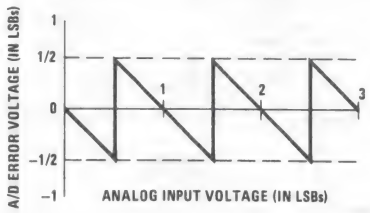
FIGURE 4. Complete Basic Tester Circuit

Typical Applications (Continued)

The digital output LED display can be decoded by dividing the 8 bits into the 4 most significant bits and 4 least significant bits. Table I shows the fractional binary equivalent of these two 8-bit groups. By adding the decoded voltages which are obtained from the column: "Input Voltage Value with a 10.240 V_{REF}" of both the MS and LS groups, the value of the digital display can be determined. For example, for an output LED display of "1011 0110" or "B6" (in hex) the voltage values from the table are 7.04 + 0.24 or 7.280 V_{DC}. These voltage values represent the center values of a perfect A/D converter. The input voltage has to change by $\pm\frac{1}{2}$ LSB ($\pm$20 mV), the "quantization uncertainty" of an A/D, to obtain an output digital code change. The effects of this quantization error have to be accounted for in the interpretation of the test results. A plot of this natural error source is shown in *Figure 5* where, for clarity, both the analog input voltage and the error voltage are normalized to LSBs.

TABLE I. DECODING THE DIGITAL OUTPUT LEDs

HEX	BINARY				FRACTIONAL BINARY VALUE FOR				INPUT VOLTAGE VALUE WITH 10.24 V_{REF}	
					MS GROUP		LS GROUP		MS GROUP	LS GROUP
F	1	1	1	1		15/16		15/256	9.600	0.600
E	1	1	1	0		7/8		7/128	8.960	0.560
D	1	1	0	1		13/16		13/256	8.320	0.520
C	1	1	0	0		3/4		3/64	7.680	0.480
B	1	0	1	1		11/16		11/256	7.040	0.440
A	1	0	1	0		5/8		5/128	6.400	0.400
9	1	0	0	1		9/16		9/256	5.760	0.360
8	1	0	0	0	1/2		1/32		5.120	0.320
7	0	1	1	1		7/16		7/256	4.480	0.280
6	0	1	1	0		3/8		3/128	3.840	0.240
5	0	1	0	1		5/16		5/256	3.200	0.200
4	0	1	0	0		1/4		1/64	2.560	0.160
3	0	0	1	1		3/16		3/256	1.920	0.120
2	0	0	1	0		1/8		1/128	1.280	0.080
1	0	0	0	1		1/16		1/256	0.640	0.040
0	0	0	0	0					0	0



TL/H/5670–8

FIGURE 5. Error Plot of a Perfect A/D Showing Effects of Quantization Error

Typical Applications (Continued)

A low speed ramp generator can also be used to sweep the analog input voltage and the LED outputs will provide a binary counting sequence from zero to full-scale.

The techniques described so far are suitable for an engineering evaluation or a quick check on performance. For a higher speed test system, or to obtain plotted data, a digital-to-analog converter is needed for the test set-up. An accurate 10-bit DAC can serve as the precision voltage source for the A/D. Errors of the A/D under test can be provided as either analog voltages or differences in two digital words.

A basic A/D tester which uses a DAC and provides the error as an analog output voltage is shown in *Figure 6*. The 2 op amps can be eliminated if a lab DVM with a numerical subtraction feature is available to directly readout the difference voltage, "A−C".

For operation with a microprocessor or a computer-based test system, it is more convenient to present the errors digitally. This can be done with the circuit of *Figure 7* where the output code transitions can be detected as the 10-bit DAC is incremented. This provides ¼ LSB steps for the 8-bit A/D under test. If the results of this test are automatically plotted with the analog input on the X axis and the error (in LSB's) as the Y axis, a useful transfer function of the A/D under test results. For acceptance testing, the plot is not necessary and the testing speed can be increased by establishing internal limits on the allowed error for each code.

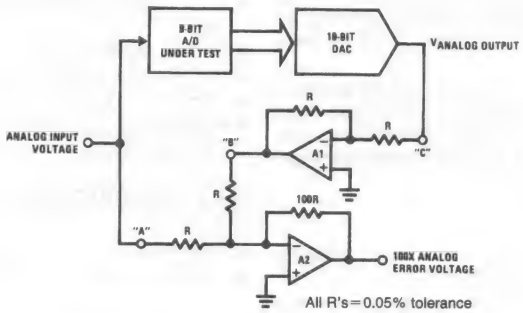


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FIGURE 6. A/D Tester with Analog Error Output

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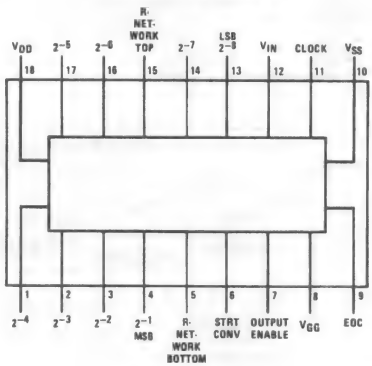
FIGURE 7. Basic "Digital" A/D Tester

Connection Diagram

Dual-In-Line Package



TL/H/5670–9

Top View

Order Number ADC0800PD
or ADC0800PCD
See NS Package Number D18A

National Semiconductor

ADC0801/ADC0802/ADC0803/ADC0804/ADC0805 8-Bit μP Compatible A/D Converters

General Description

The ADC0801, ADC0802, ADC0803, ADC0804 and ADC0805 are CMOS 8-bit successive approximation A/D converters that use a differential potentiometric ladder—similar to the 256R products. These converters are designed to allow operation with the NSC800 and INS8080A derivative control bus with TRI-STATE® output latches directly driving the data bus. These A/Ds appear like memory locations or I/O ports to the microprocessor and no interfacing logic is needed.

Differential analog voltage inputs allow increasing the common-mode rejection and offsetting the analog zero input voltage value. In addition, the voltage reference input can be adjusted to allow encoding any smaller analog voltage span to the full 8 bits of resolution.

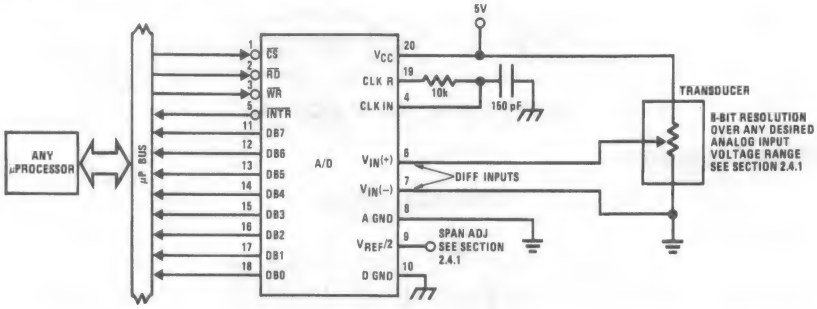
Features

- Compatible with 8080 μP derivatives—no interfacing logic needed - access time - 135 ns
- Easy interface to all microprocessors, or operates "stand alone"
- Differential analog voltage inputs
- Logic inputs and outputs meet both MOS and TTL voltage level specifications
- Works with 2.5V (LM336) voltage reference
- On-chip clock generator
- 0V to 5V analog input voltage range with single 5V supply
- No zero adjust required
- 0.3" standard width 20-pin DIP package
- 20-pin molded chip carrier or small outline package
- Operates ratiometrically or with 5 V_{DC}, 2.5 V_{DC}, or analog span adjusted voltage reference

Key Specifications

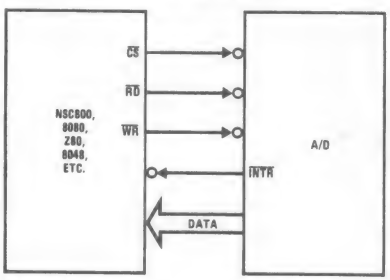
- Resolution: 8 bits
- Total error: ±¼ LSB, ±½ LSB and ±1 LSB
- Conversion time: 100 μs

Typical Applications



TL/H/5671–1

8080 Interface



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Error Specification (Includes Full-Scale, Zero Error, and Non-Linearity)			
Part Number	Full-Scale Adjusted	$V_{REF}/2 = 2.500\ V_{DC}$ (No Adjustments)	$V_{REF}/2$ = No Connection (No Adjustments)
ADC0801	±¼ LSB		
ADC0802		±½ LSB	
ADC0803	±½ LSB		
ADC0804		±1 LSB	
ADC0805			±1 LSB

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC}) (Note 3)	6.5V
Voltage	
Logic Control Inputs	−0.3V to +18V
At Other Input and Outputs	−0.3V to (V_{CC} + 0.3V)
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
Surface Mount Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C
Storage Temperature Range	−65°C to +150°C
Package Dissipation at T_A = 25°C	875 mW
ESD Susceptibility (Note 10)	800V

Operating Ratings (Notes 1 & 2)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0801/02LJ	$-55°C \le T_A \le +125°C$
ADC0801/02/03/04LCJ	$-40°C \le T_A \le +85°C$
ADC0801/02/03/05LCN	$-40°C \le T_A \le +85°C$
ADC0804LCN	$0°C \le T_A \le +70°C$
ADC0802/03/04LCV	$0°C \le T_A \le +70°C$
ADC0802/03/04LCWM	$0°C \le T_A \le +70°C$
Range of V_{CC}	4.5 V_{DC} to 6.3 V_{DC}

Electrical Characteristics

The following specifications apply for $V_{CC} = 5\ V_{DC}$, $T_{MIN} \le T_A \le T_{MAX}$ and $f_{CLK} = 640$ kHz unless otherwise specified.

Parameter	Conditions	Min	Typ	Max	Units
ADC0801: Total Adjusted Error (Note 8)	With Full-Scale Adj. (See Section 2.5.2)			±¼	LSB
ADC0802: Total Unadjusted Error (Note 8)	$V_{REF}/2 = 2.500\ V_{DC}$			±½	LSB
ADC0803: Total Adjusted Error (Note 8)	With Full-Scale Adj. (See Section 2.5.2)			±½	LSB
ADC0804: Total Unadjusted Error (Note 8)	$V_{REF}/2 = 2.500\ V_{DC}$			±1	LSB
ADC0805: Total Unadjusted Error (Note 8)	$V_{REF}/2$-No Connection			±1	LSB
$V_{REF}/2$ Input Resistance (Pin 9)	ADC0801/02/03/05 ADC0804 (Note 9)	2.5 0.75	8.0 1.1		kΩ kΩ
Analog Input Voltage Range	(Note 4) V(+) or V(−)	Gnd−0.05		V_{CC} + 0.05	V_{DC}
DC Common-Mode Error	Over Analog Input Voltage Range		±1/16	±⅛	LSB
Power Supply Sensitivity	$V_{CC} = 5\ V_{DC}$ ±10% Over Allowed $V_{IN}(+)$ and $V_{IN}(-)$ Voltage Range (Note 4)		±1/16	±⅛	LSB

AC Electrical Characteristics

The following specifications apply for $V_{CC} = 5\ V_{DC}$ and $T_A = 25°C$ unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
T_C	Conversion Time	$f_{CLK} = 640$ kHz (Note 6)	103		114	μs
T_C	Conversion Time	(Note 5, 6)	66		73	$1/f_{CLK}$
f_{CLK}	Clock Frequency Clock Duty Cycle	$V_{CC} = 5V$, (Note 5) (Note 5)	100 40	640	1460 60	kHz %
CR	Conversion Rate in Free-Running Mode	$\overline{INTR}$ tied to $\overline{WR}$ with $\overline{CS} = 0\ V_{DC}$, $f_{CLK} = 640$ kHz	8770		9708	conv/s
$t_{W(\overline{WR})L}$	Width of $\overline{WR}$ Input (Start Pulse Width)	$\overline{CS} = 0\ V_{DC}$ (Note 7)	100			ns
t_{ACC}	Access Time (Delay from Falling Edge of $\overline{RD}$ to Output Data Valid)	$C_L = 100$ pF		135	200	ns
t_{1H}, t_{0H}	TRI-STATE Control (Delay from Rising Edge of $\overline{RD}$ to Hi-Z State)	$C_L = 10$ pF, $R_L = 10$k (See TRI-STATE Test Circuits)		125	200	ns
t_{WI}, t_{RI}	Delay from Falling Edge of $\overline{WR}$ or $\overline{RD}$ to Reset of $\overline{INTR}$			300	450	ns
C_{IN}	Input Capacitance of Logic Control Inputs			5	7.5	pF
C_{OUT}	TRI-STATE Output Capacitance (Data Buffers)			5	7.5	pF
CONTROL INPUTS [Note: CLK IN (Pin 4) is the input of a Schmitt trigger circuit and is therefore specified separately]						
V_{IN} (1)	Logical "1" Input Voltage (Except Pin 4 CLK IN)	$V_{CC} = 5.25\ V_{DC}$	2.0		15	V_{DC}

AC Electrical Characteristics (Continued)

The following specifications apply for $V_{CC} = 5V_{DC}$ and $T_{MIN} \leq T_A \leq T_{MAX}$, unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
CONTROL INPUTS [Note: CLK IN (Pin 4) is the input of a Schmitt trigger circuit and is therefore specified separately]						
V_{IN} (0)	Logical "0" Input Voltage (Except Pin 4 CLK IN)	$V_{CC} = 4.75\ V_{DC}$			0.8	V_{DC}
I_{IN} (1)	Logical "1" Input Current (All Inputs)	$V_{IN} = 5\ V_{DC}$		0.005	1	μA_{DC}
I_{IN} (0)	Logical "0" Input Current (All Inputs)	$V_{IN} = 0\ V_{DC}$	−1	−0.005		μA_{DC}
CLOCK IN AND CLOCK R						
V_T+	CLK IN (Pin 4) Positive Going Threshold Voltage		2.7	3.1	3.5	V_{DC}
V_T-	CLK IN (Pin 4) Negative Going Threshold Voltage		1.5	1.8	2.1	V_{DC}
V_H	CLK IN (Pin 4) Hysteresis $(V_T+)-(V_T-)$		0.6	1.3	2.0	V_{DC}
V_{OUT} (0)	Logical "0" CLK R Output Voltage	$I_O = 360\ \mu A$ $V_{CC} = 4.75\ V_{DC}$			0.4	V_{DC}
V_{OUT} (1)	Logical "1" CLK R Output Voltage	$I_O = -360\ \mu A$ $V_{CC} = 4.75\ V_{DC}$	2.4			V_{DC}
DATA OUTPUTS AND $\overline{INTR}$						
V_{OUT} (0)	Logical "0" Output Voltage Data Outputs $\overline{INTR}$ Output	 $I_{OUT} = 1.6$ mA, $V_{CC} = 4.75\ V_{DC}$ $I_{OUT} = 1.0$ mA, $V_{CC} = 4.75\ V_{DC}$			 0.4 0.4	 V_{DC} V_{DC}
V_{OUT} (1)	Logical "1" Output Voltage	$I_O = -360\ \mu A$, $V_{CC} = 4.75\ V_{DC}$	2.4			V_{DC}
V_{OUT} (1)	Logical "1" Output Voltage	$I_O = -10\ \mu A$, $V_{CC} = 4.75\ V_{DC}$	4.5			V_{DC}
I_{OUT}	TRI-STATE Disabled Output Leakage (All Data Buffers)	$V_{OUT} = 0\ V_{DC}$ $V_{OUT} = 5\ V_{DC}$	−3		 3	μA_{DC} μA_{DC}
I_{SOURCE}		V_{OUT} Short to Gnd, $T_A = 25°C$	4.5	6		mA_{DC}
I_{SINK}		V_{OUT} Short to V_{CC}, $T_A = 25°C$	9.0	16		mA_{DC}
POWER SUPPLY						
I_{CC}	Supply Current (Includes Ladder Current) ADC0801/02/03/04LCJ/05 ADC0804LCN/LCV/LCWM	$f_{CLK} = 640$ kHz, $V_{REF}/2 =$ NC, $T_A = 25°C$ and $\overline{CS} = 5V$		 1.1 1.9	 1.8 2.5	 mA mA

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to Gnd, unless otherwise specified. The separate A Gnd point should always be wired to the D Gnd.

Note 3: A zener diode exists, internally, from V_{CC} to Gnd and has a typical breakdown voltage of 7 V_{DC}.

Note 4: For $V_{IN}(-) \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input (see block diagram) which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct–especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

Note 5: Accuracy is guaranteed at $f_{CLK} = 640$ kHz. At higher clock frequencies accuracy can degrade. For lower clock frequencies, the duty cycle limits can be extended so long as the minimum clock high time interval or minimum clock low time interval is no less than 275 ns.

Note 6: With an asynchronous start pulse, up to 8 clock periods may be required before the internal clock phases are proper to start the conversion process. The start request is internally latched, see *Figure 2* and section 2.0.

Note 7: The $\overline{CS}$ input is assumed to bracket the $\overline{WR}$ strobe input and therefore timing is dependent on the $\overline{WR}$ pulse width. An arbitrarily wide pulse width will hold the converter in a reset mode and the start of conversion is initiated by the low to high transition of the $\overline{WR}$ pulse (see timing diagrams).

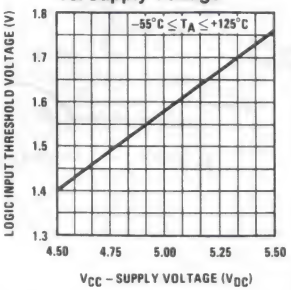
Note 8: None of these A/Ds requires a zero adjust (see section 2.5.1). To obtain zero code at other analog input voltages see section 2.5 and *Figure 5*.

Note 9: The $V_{REF}/2$ pin is the center point of a two-resistor divider connected from V_{CC} to ground. In all versions of the ADC0801, ADC0802, ADC0803, and ADC0805, and in the ADC0804LCJ, each resistor is typically 16 kΩ. In all versions of the ADC0804 except the ADC0804LCJ, each resistor is typically 2.2 kΩ.

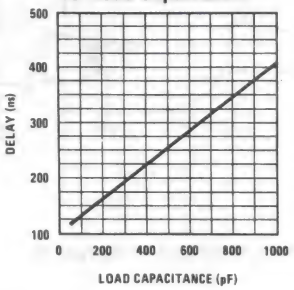
Note 10: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Typical Performance Characteristics

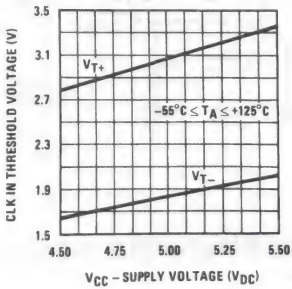
Logic Input Threshold Voltage vs. Supply Voltage



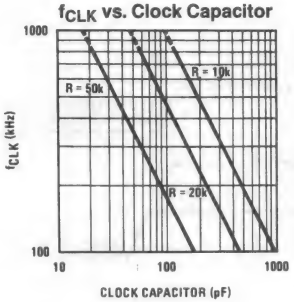
Delay From Falling Edge of $\overline{RD}$ to Output Data Valid vs. Load Capacitance



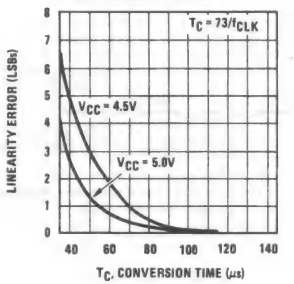
CLK IN Schmitt Trip Levels vs. Supply Voltage



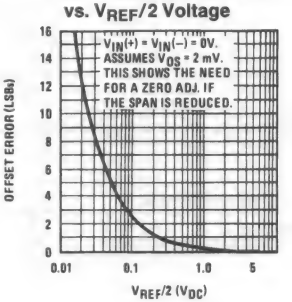
f_{CLK} vs. Clock Capacitor



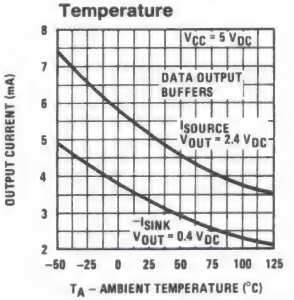
Full-Scale Error vs Conversion Time



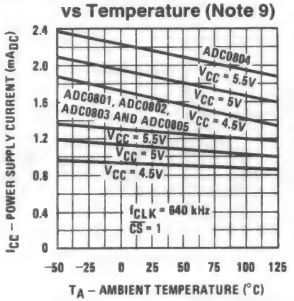
Effect of Unadjusted Offset Error vs. $V_{REF}/2$ Voltage



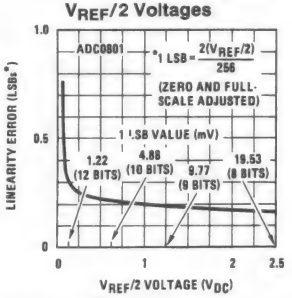
Output Current vs Temperature



Power Supply Current vs Temperature (Note 9)



Linearity Error at Low $V_{REF}/2$ Voltages



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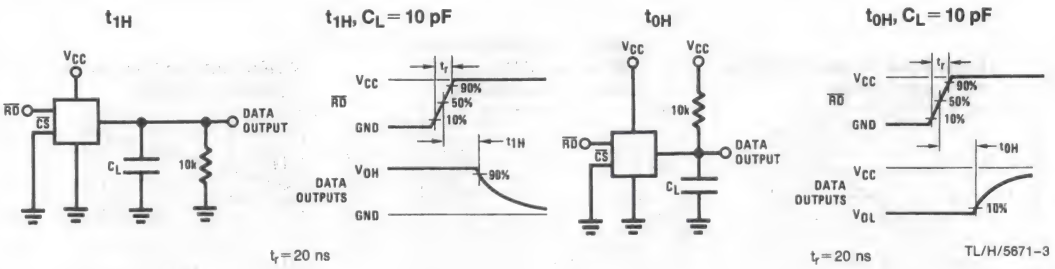
TRI-STATE Test Circuits and Waveforms



t_r = 20 ns

t_r = 20 ns

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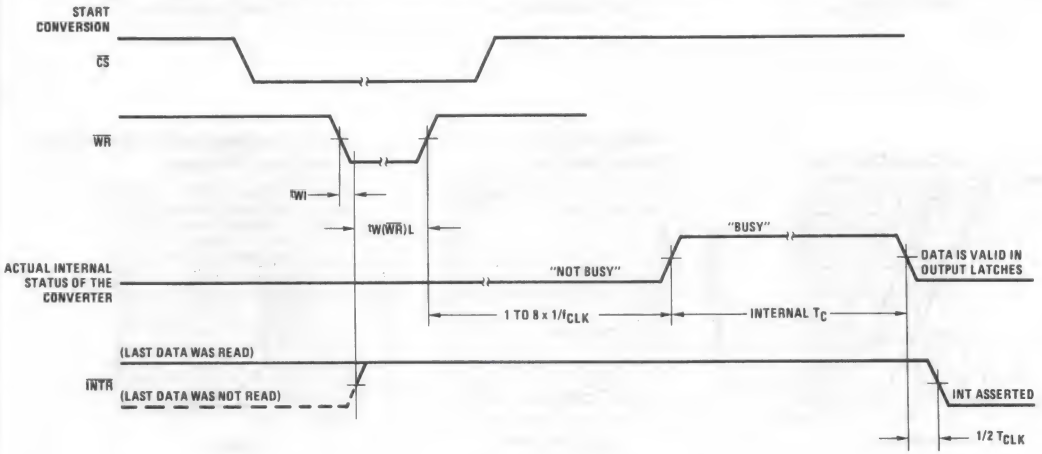
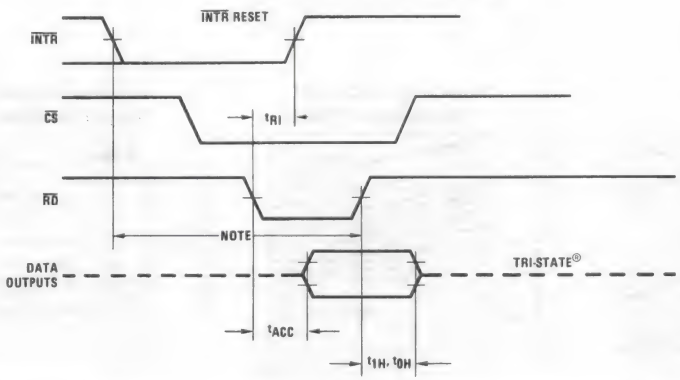
Timing Diagrams (All timing is measured from the 50% voltage points)



Output Enable and Reset $\overline{INTR}$



Note: Read strobe must occur 8 clock periods (8/f_{CLK}) after assertion of interrupt to guarantee reset of $\overline{INTR}$.

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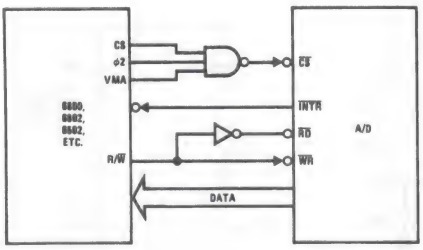
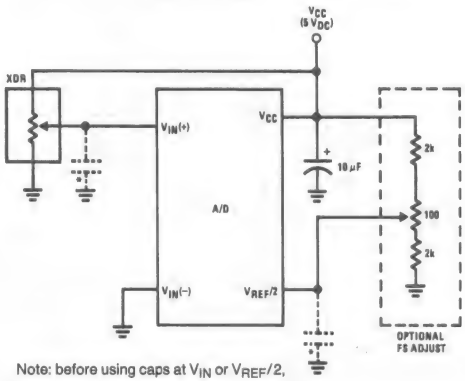
Typical Applications (Continued)

6800 Interface

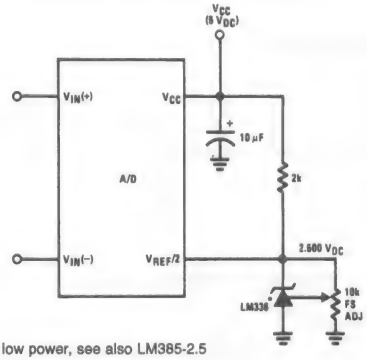


Ratiometric with Full-Scale Adjust



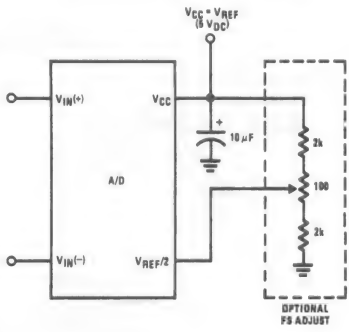
Note: before using caps at V_{IN} or $V_{REF}/2$, see section 2.3.2 Input Bypass Capacitors.

Absolute with a 2.500V Reference

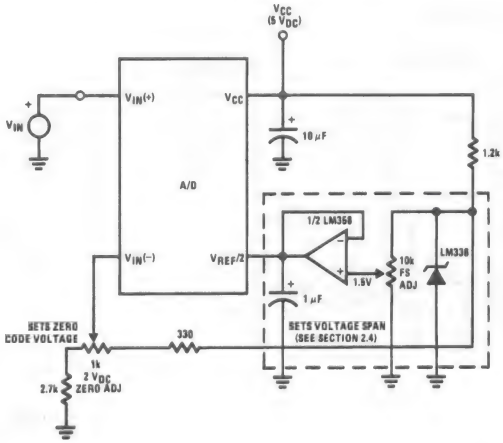


*For low power, see also LM385-2.5

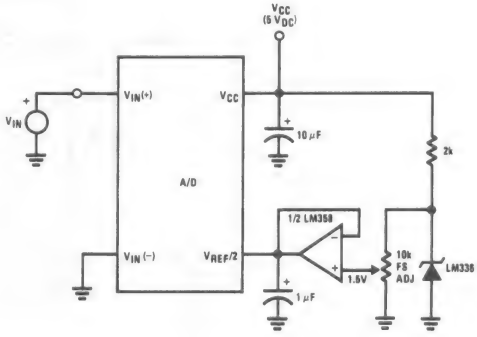
Absolute with a 5V Reference



Zero-Shift and Span Adjust: $2V \leq V_{IN} \leq 5V$



Span Adjust: $0V \leq V_{IN} \leq 3V$



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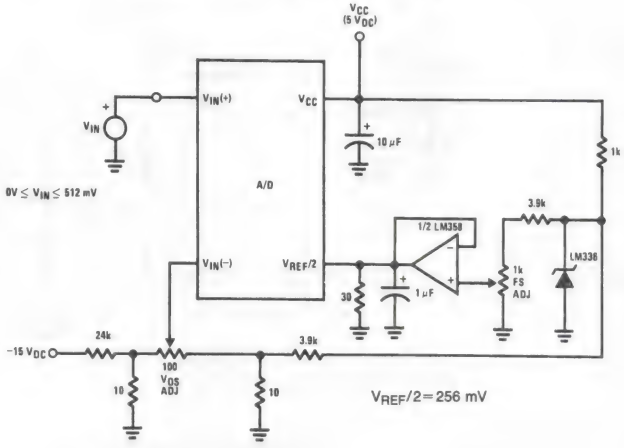
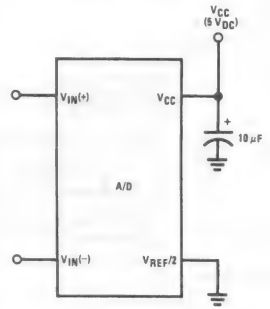
Typical Applications (Continued)

Directly Converting a Low-Level Signal



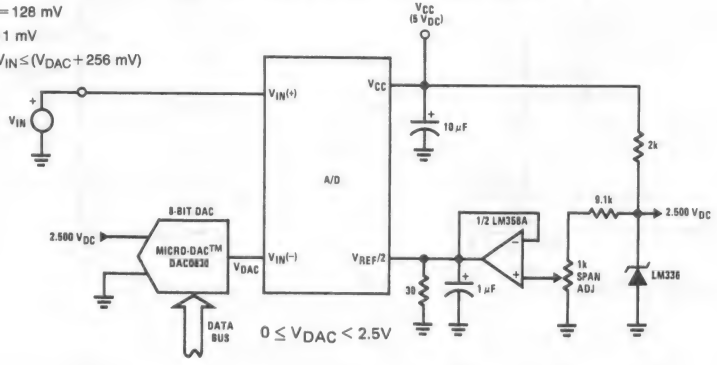
A μP Interfaced Comparator



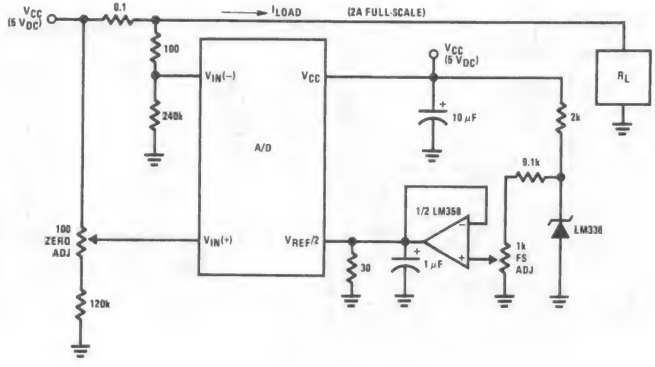
For: $V_{IN}(+) > V_{IN}(-)$
Output = FF_{HEX}
For: $V_{IN}(+) < V_{IN}(-)$
Output = 00_{HEX}

1 mV Resolution with μP Controlled Range

$V_{REF}/2$ = 128 mV
1 LSB = 1 mV
$V_{DAC} \leq V_{IN} \leq (V_{DAC} + 256\ mV)$



Digitizing a Current Flow



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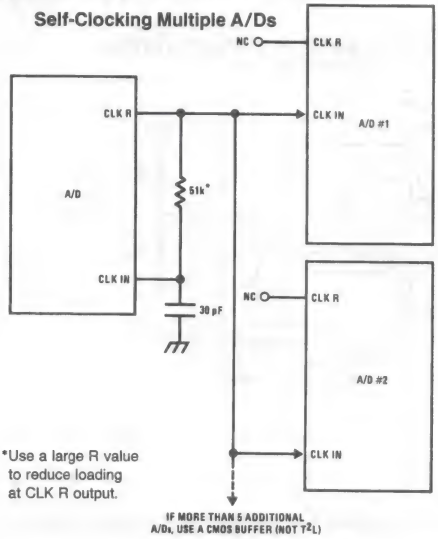
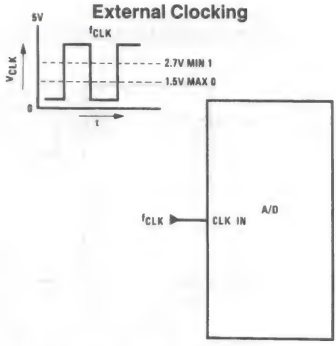
Typical Applications (Continued)

Self-Clocking Multiple A/Ds



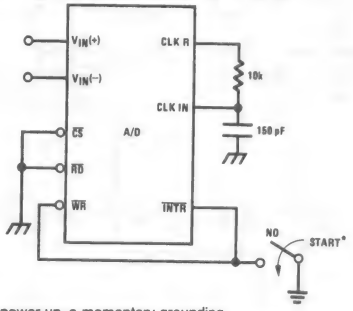
*Use a large R value to reduce loading at CLK R output.

External Clocking



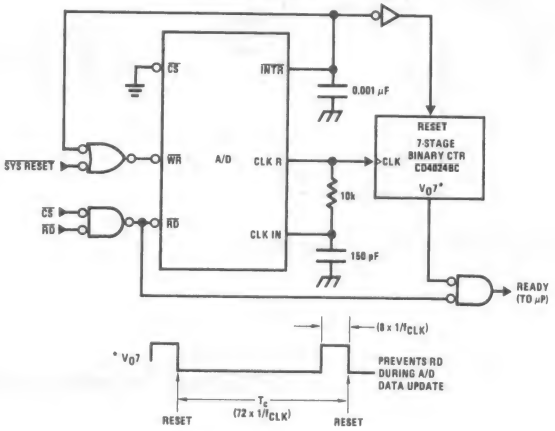
100 kHz ≤ f_{CLK} ≤ 1460 kHz

Self-Clocking in Free-Running Mode

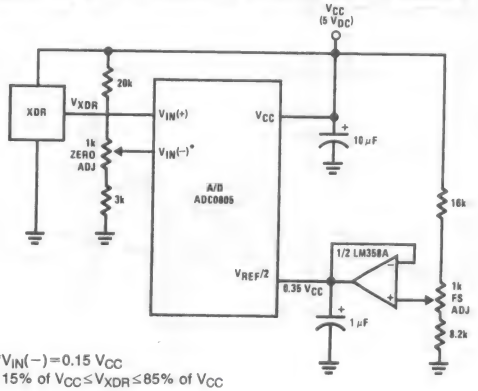


*After power-up, a momentary grounding of the $\overline{WR}$ input is needed to guarantee operation.

μP Interface for Free-Running A/D

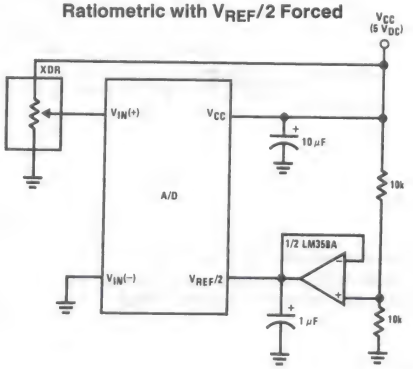


Operating with "Automotive" Ratiometric Transducers



*$V_{IN}(-) = 0.15\ V_{CC}$
15% of $V_{CC} \leq V_{XDR} \leq 85\%$ of V_{CC}

Ratiometric with $V_{REF}/2$ Forced



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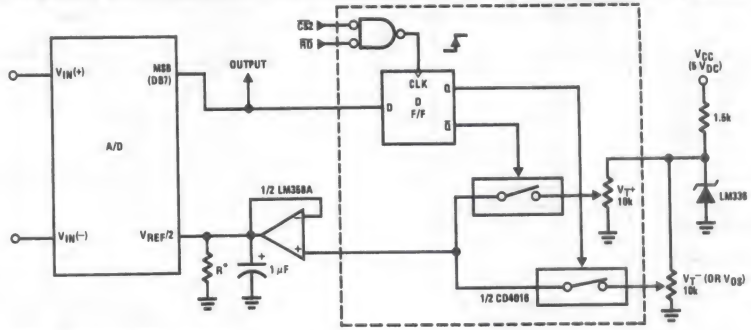
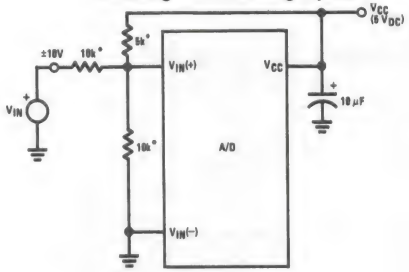
Typical Applications (Continued)

μP Compatible Differential-Input Comparator with Pre-Set V_{OS} (with or without Hysteresis)



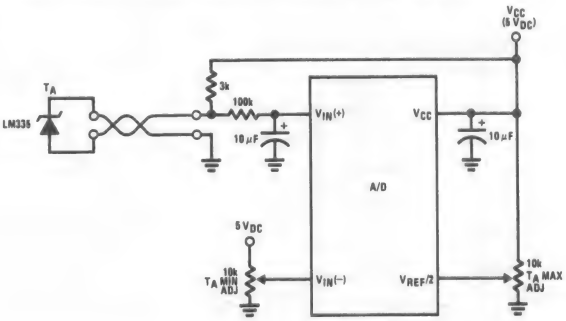
*See *Figure 5* to select R value
DB7 = "1" for $V_{IN}(+) > V_{IN}(-) + (V_{REF}/2)$
Omit circuitry within the dotted area if hysteresis is not needed

Handling ±10V Analog Inputs

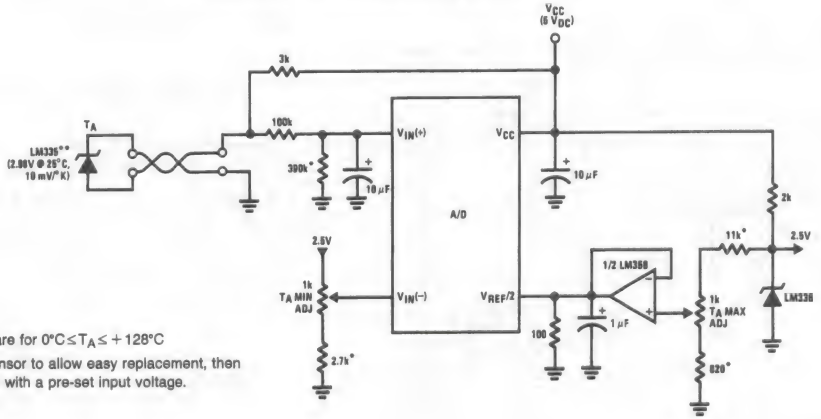


*Beckman Instruments #694-3-R10K resistor array

Low-Cost, μP Interfaced, Temperature-to-Digital Converter



μP Interfaced Temperature-to-Digital Converter



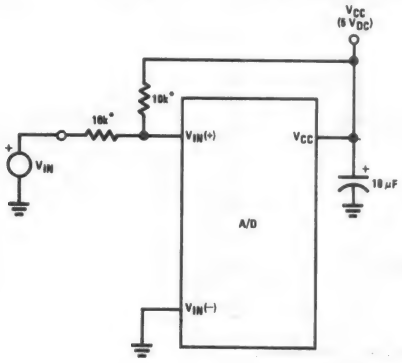
*Circuit values shown are for $0°C \leq T_A \leq +128°C$

**Can calibrate each sensor to allow easy replacement, then A/D can be calibrated with a pre-set input voltage.

TL/H/5671–8

Typical Applications (Continued)

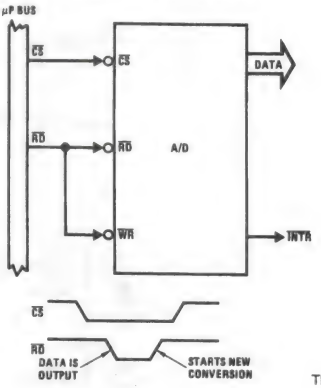
Handling ±5V Analog Inputs



TL/H/5671–33

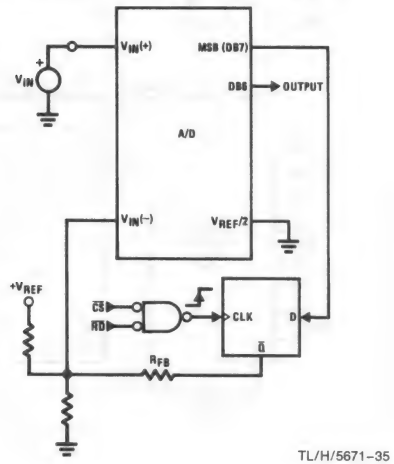
*Beckman Instruments #694-3-R10K resistor array

Read-Only Interface



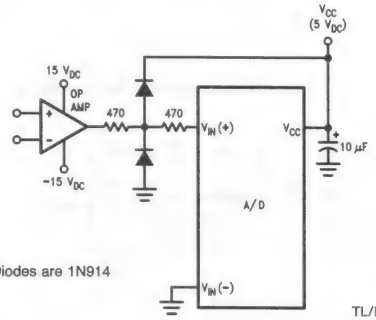
TL/H/5671–34

µP Interfaced Comparator with Hysteresis



TL/H/5671–35

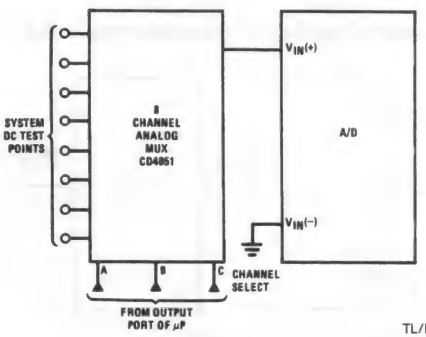
Protecting the Input



Diodes are 1N914

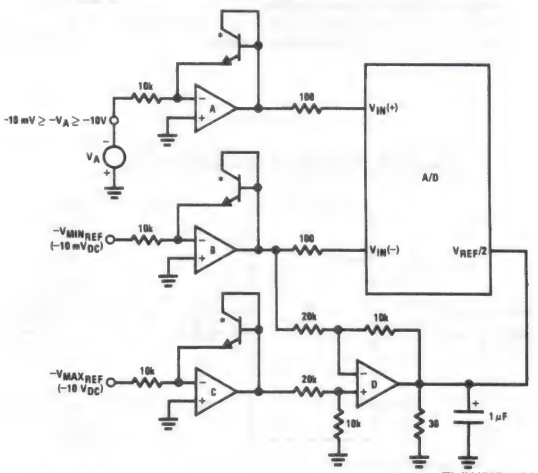
TL/H/5671–9

Analog Self-Test for a System



TL/H/5671–36

A Low-Cost, 3-Decade Logarithmic Converter



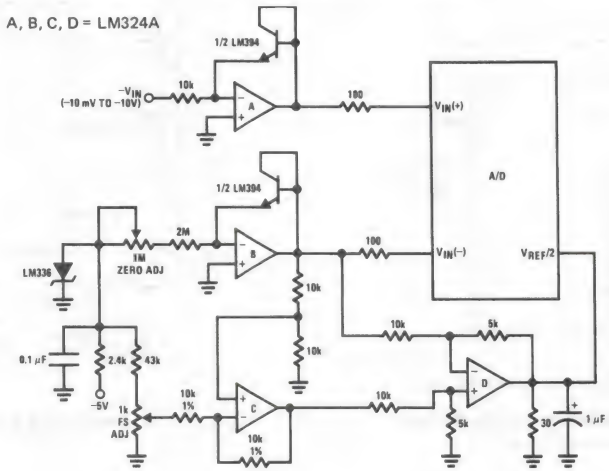
TL/H/5671–37

*LM389 transistors

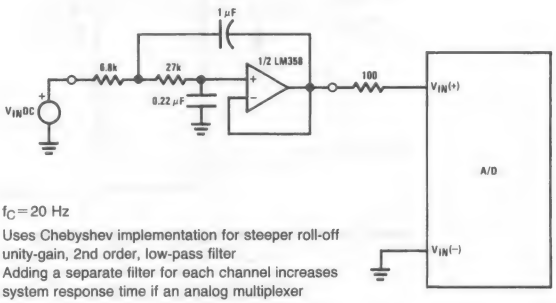
A, B, C, D = LM324A quad op amp

Typical Applications (Continued)

3-Decade Logarithmic A/D Converter



Noise Filtering the Analog Input

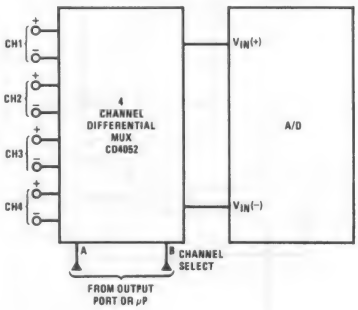


$f_C = 20$ Hz

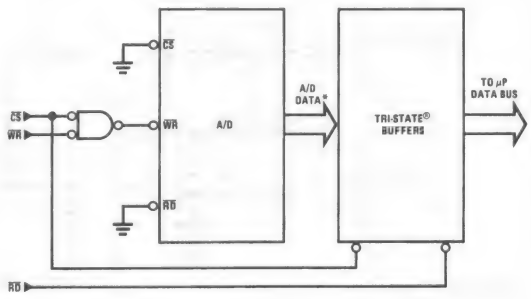
Uses Chebyshev implementation for steeper roll-off unity-gain, 2nd order, low-pass filter

Adding a separate filter for each channel increases system response time if an analog multiplexer is used

Multiplexing Differential Inputs

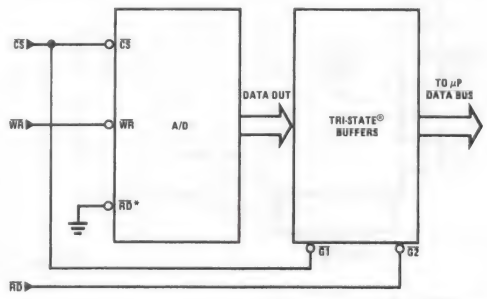


Output Buffers with A/D Data Enabled



*A/D output data is updated 1 CLK period prior to assertion of $\overline{INTR}$

Increasing Bus Drive and/or Reducing Time on Bus

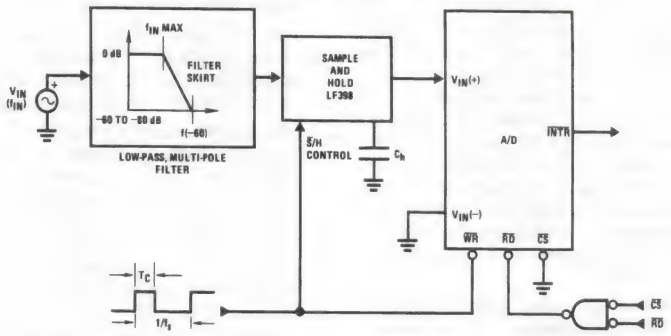


TL/H/5671–10

*Allows output data to set-up at falling edge of $\overline{CS}$

Typical Applications (Continued)

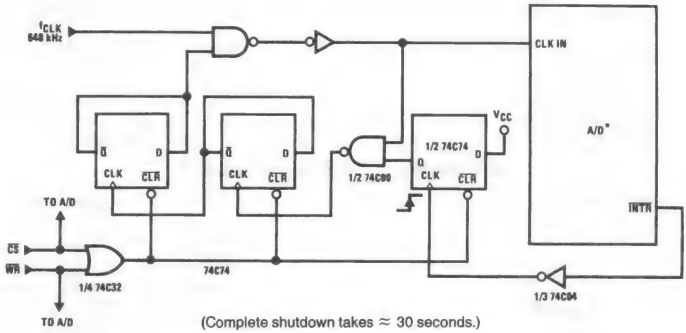
Sampling an AC Input Signal



Note 1: Oversample whenever possible [keep fs > 2f(−60)] to eliminate input frequency folding (aliasing) and to allow for the skirt response of the filter.

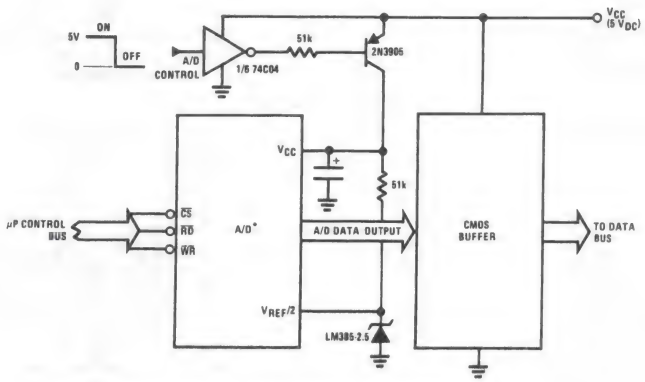
Note 2: Consider the amplitude errors which are introduced within the passband of the filter.

70% Power Savings by Clock Gating



(Complete shutdown takes ≈ 30 seconds.)

Power Savings by A/D and V_{REF} Shutdown



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*Use ADC0801, 02, 03 or 05 for lowest power consumption.

Note: Logic inputs can be driven to V_{CC} with A/D supply at zero volts.

Buffer prevents data bus from overdriving output of A/D when in shutdown mode.

Functional Description

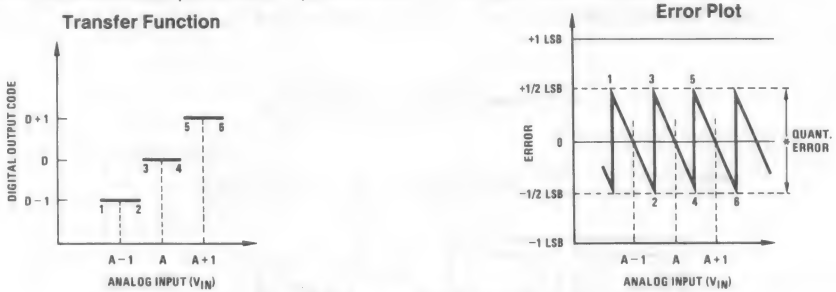
1.0 UNDERSTANDING A/D ERROR SPECS

A perfect A/D transfer characteristic (staircase waveform) is shown in *Figure 1a*. The horizontal scale is analog input voltage and the particular points labeled are in steps of 1 LSB (19.53 mV with 2.5V tied to the $V_{REF}/2$ pin). The digital output codes that correspond to these inputs are shown as D−1, D, and D+1. For the perfect A/D, not only will center-value (A−1, A, A+1, ...) analog inputs produce the correct output ditigal codes, but also each riser (the transitions between adjacent output codes) will be located ±½ LSB away from each center-value. As shown, the risers are ideal and have no width. Correct digital output codes will be provided for a range of analog input voltages that extend ±½ LSB from the ideal center-values. Each tread (the range of analog input voltage that provides the same digital output code) is therefore 1 LSB wide.

Figure 1b shows a worst case error plot for the ADC0801. All center-valued inputs are guaranteed to produce the correct output codes and the adjacent risers are guaranteed to be no closer to the center-value points than ±¼ LSB. In other words, if we apply an analog input equal to the center-value ±¼ LSB, *we guarantee* that the A/D will produce the correct digital code. The maximum range of the position of the code transition is indicated by the horizontal arrow and it is guaranteed to be no more than ½ LSB.

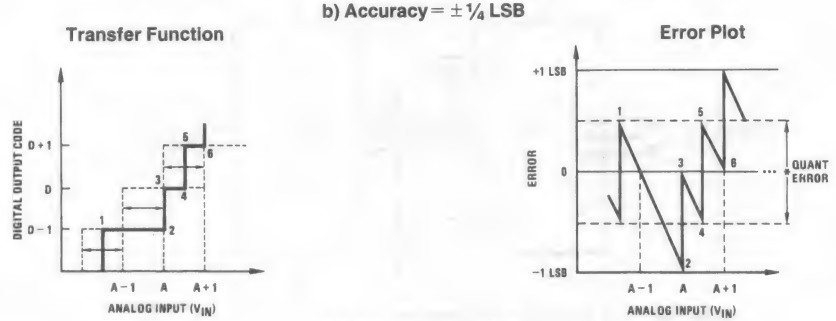
The error curve of *Figure 1c* shows a worst case error plot for the ADC0802. Here we guarantee that if we apply an analog input equal to the LSB analog voltage center-value the A/D will produce the correct digital code.

Next to each transfer function is shown the corresponding error plot. Many people may be more familiar with error plots than transfer functions. The analog input voltage to the A/D is provided by either a linear ramp or by the discrete output steps of a high resolution DAC. Notice that the error is continuously displayed and includes the quantization uncertainty of the A/D. For example the error at point 1 of *Figure 1a* is +½ LSB because the digital code appeared ½ LSB in advance of the center-value of the tread. The error plots always have a constant negative slope and the abrupt upside steps are always 1 LSB in magnitude.



a) Accuracy = ±0 LSB: A Perfect A/D

b) Accuracy = ±¼ LSB



c) Accuracy = ±½ LSB

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FIGURE 1. Clarifying the Error Specs of an A/D Converter

Functional Description (Continued)

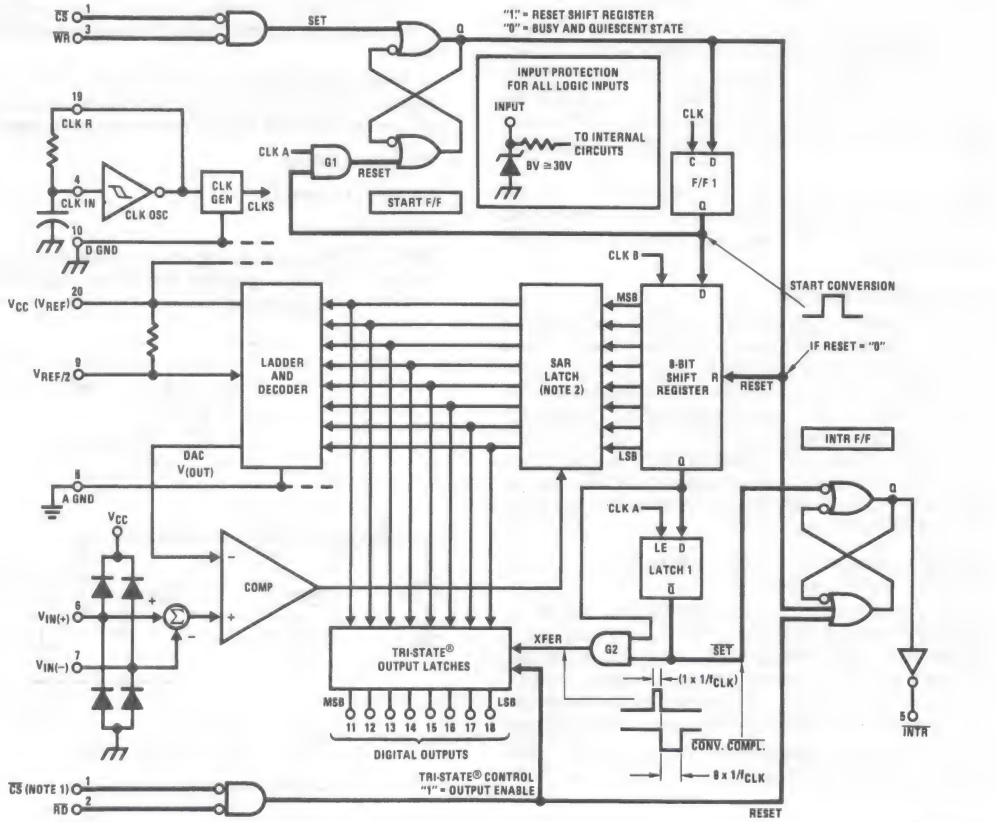
2.0 FUNCTIONAL DESCRIPTION

The ADC0801 series contains a circuit equivalent of the 256R network. Analog switches are sequenced by successive approximation logic to match the analog difference input voltage $[V_{IN}(+) - V_{IN}(-)]$ to a corresponding tap on the R network. The most significant bit is tested first and after 8 comparisons (64 clock cycles) a digital 8-bit binary code (1111 1111 = full-scale) is transferred to an output latch and then an interrupt is asserted ($\overline{INTR}$ makes a high-to-low transition). A conversion in process can be interrupted by issuing a second start command. The device may be operated in the free-running mode by connecting $\overline{INTR}$ to the $\overline{WR}$ input with $\overline{CS}=0$. To ensure start-up under all possible conditions, an external $\overline{WR}$ pulse is required during the first power-up cycle.

On the high-to-low transition of the $\overline{WR}$ input the internal SAR latches and the shift register stages are reset. As long as the $\overline{CS}$ input and $\overline{WR}$ input remain low, the A/D will remain in a reset state. *Conversion will start from 1 to 8 clock periods after at least one of these inputs makes a low-to-high transition.*

A functional diagram of the A/D converter is shown in *Figure 2*. All of the package pinouts are shown and the major logic control paths are drawn in heavier weight lines.

The converter is started by having $\overline{CS}$ and $\overline{WR}$ simultaneously low. This sets the start flip-flop (F/F) and the resulting "1" level resets the 8-bit shift register, resets the Interrupt (INTR) F/F and inputs a "1" to the D flop, F/F1, which is at the input end of the 8-bit shift register. Internal clock signals then transfer this "1" to the Q output of F/F1. The AND gate, G1, combines this "1" output with a clock signal to provide a reset signal to the start F/F. If the set signal is no longer present (either $\overline{WR}$ or $\overline{CS}$ is a "1") the start F/F is reset and the 8-bit shift register then can have the "1" clocked in, which starts the conversion process. If the set signal were to still be present, this reset pulse would have no effect (both outputs of the start F/F would momentarily be at a "1" level) and the 8-bit shift register would continue to be held in the reset mode. This logic therefore allows for wide $\overline{CS}$ and $\overline{WR}$ signals and the converter will start after at least one of these signals returns high and the internal clocks again provide a reset signal for the start F/F.



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Note 1: $\overline{CS}$ shown twice for clarity.

Note 2: SAR = Successive Approximation Register.

FIGURE 2. Block Diagram

Functional Description (Continued)

After the "1" is clocked through the 8-bit shift register (which completes the SAR search) it appears as the input to the D-type latch, LATCH 1. As soon as this "1" is output from the shift register, the AND gate, G2, causes the new digital word to transfer to the TRI-STATE output latches. When LATCH 1 is subsequently enabled, the Q output makes a high-to-low transition which causes the INTR F/F to set. An inverting buffer then supplies the $\overline{INTR}$ input signal.

Note that this $\overline{SET}$ control of the INTR F/F remains low for 8 of the external clock periods (as the internal clocks run at 1/8 of the frequency of the external clock). If the data output is continuously enabled ($\overline{CS}$ and $\overline{RD}$ both held low), the $\overline{INTR}$ output will still signal the end of conversion (by a high-to-low transition), because the $\overline{SET}$ input can control the Q output of the INTR F/F even though the RESET input is constantly at a "1" level in this operating mode. This $\overline{INTR}$ output will therefore stay low for the duration of the $\overline{SET}$ signal, which is 8 periods of the external clock frequency (assuming the A/D is not started during this interval).

When operating in the free-running or continuous conversion mode ($\overline{INTR}$ pin tied to $\overline{WR}$ and $\overline{CS}$ wired low—see also section 2.8), the START F/F is SET by the high-to-low transition of the $\overline{INTR}$ signal. This resets the SHIFT REGISTER which causes the input to the D-type latch, LATCH 1, to go low. As the latch enable input is still present, the $\overline{Q}$ output will go high, which then allows the INTR F/F to be RESET. This reduces the width of the resulting $\overline{INTR}$ output pulse to only a few propagation delays (approximately 300 ns).

When data is to be read, the combination of both $\overline{CS}$ and $\overline{RD}$ being low will cause the INTR F/F to be reset and the TRI-STATE output latches will be enabled to provide the 8-bit digital outputs.

2.1 Digital Control Inputs

The digital control inputs ($\overline{CS}$, $\overline{RD}$, and $\overline{WR}$) meet standard T^{2}L logic voltage levels. These signals have been renamed when compared to the standard A/D Start and Output Enable labels. In addition, these inputs are active low to allow an easy interface to microprocessor control busses. For non-microprocessor based applications, the $\overline{CS}$ input (pin 1) can be grounded and the standard A/D Start function is obtained by an active low pulse applied at the $\overline{WR}$ input (pin 3) and the Output Enable function is caused by an active low pulse at the $\overline{RD}$ input (pin 2).

2.2 Analog Differential Voltage Inputs and Common-Mode Rejection

This A/D has additional applications flexibility due to the analog differential voltage input. The $V_{IN}(-)$ input (pin 7) can be used to automatically subtract a fixed voltage value from the input reading (tare correction). This is also useful in 4 mA–20 mA current loop conversion. In addition, common-mode noise can be reduced by use of the differential input.

The time interval between sampling $V_{IN}(+)$ and $V_{IN}(-)$ is 4-1/2 clock periods. The maximum error voltage due to this slight time difference between the input voltage samples is given by:

$$\Delta V_e(MAX) = (V_P)(2\pi f_{cm})\left(\frac{4.5}{f_{CLK}}\right),$$

where:

ΔV_e is the error voltage due to sampling delay

V_P is the peak value of the common-mode voltage

f_{cm} is the common-mode frequency

As an example, to keep this error to 1/4 LSB (~5 mV) when operating with a 60 Hz common-mode frequency, f_{cm}, and using a 640 kHz A/D clock, f_{CLK}, would allow a peak value of the common-mode voltage, V_P, which is given by:

$$V_P = \frac{[\Delta V_{e(MAX)}\,(f_{CLK})]}{(2\pi f_{cm})\,(4.5)}$$

or

$$V_P = \frac{(5 \times 10^{-3})\,(640 \times 10^3)}{(6.28)\,(60)\,(4.5)}$$

which gives

$V_P \cong 1.9V$.

The allowed range of analog input voltages usually places more severe restrictions on input common-mode noise levels.

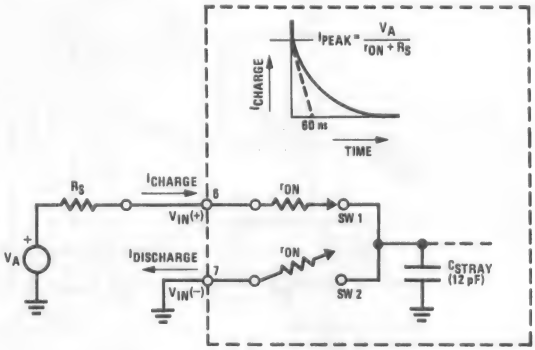
An analog input voltage with a reduced span and a relatively large zero offset can be handled easily by making use of the differential input (see section 2.4 Reference Voltage).

2.3 Analog Inputs

2.3.1 Input Current

Normal Mode

Due to the internal switching action, displacement currents will flow at the analog inputs. This is due to on-chip stray capacitance to ground as shown in *Figure 3*.



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r_{ON} of SW 1 and SW 2 ≅ 5 kΩ

$r = r_{ON}\,C_{STRAY} \cong 5\ k\Omega \times 12\ pF = 60\ ns$

FIGURE 3. Analog Input Impedance

Functional Description (Continued)

The voltage on this capacitance is switched and will result in currents entering the $V_{IN}(+)$ input pin and leaving the $V_{IN}(-)$ input which will depend on the analog differential input voltage levels. These current transients occur at the leading edge of the internal clocks. They rapidly decay and *do not cause errors* as the on-chip comparator is strobed at the end of the clock period.

Fault Mode

If the voltage source applied to the $V_{IN}(+)$ or $V_{IN}(-)$ pin exceeds the allowed operating range of $V_{CC}+50$ mV, large input currents can flow through a parasitic diode to the V_{CC} pin. If these currents can exceed the 1 mA max allowed spec, an external diode (1N914) should be added to bypass this current to the V_{CC} pin (with the current bypassed with this diode, the voltage at the $V_{IN}(+)$ pin can exceed the V_{CC} voltage by the forward voltage of this diode).

2.3.2 Input Bypass Capacitors

Bypass capacitors at the inputs will average these charges and cause a DC current to flow through the output resistances of the analog signal sources. This charge pumping action is worse for continuous conversions with the $V_{IN}(+)$ input voltage at full-scale. For continuous conversions with a 640 kHz clock frequency with the $V_{IN}(+)$ input at 5V, this DC current is at a maximum of approximately 5 μA. Therefore, *bypass capacitors should not be used at the analog inputs or the $V_{REF}/2$ pin* for high resistance sources (> 1 kΩ). If input bypass capacitors are necessary for noise filtering and high source resistance is desirable to minimize capacitor size, the detrimental effects of the voltage drop across this input resistance, which is due to the average value of the input current, can be eliminated with a full-scale adjustment while the given source resistor and input bypass capacitor are both in place. This is possible because the average value of the input current is a precise linear function of the differential input voltage.

2.3.3 Input Source Resistance

Large values of source resistance where an input bypass capacitor is not used, *will not cause errors* as the input currents settle out prior to the comparison time. If a low pass filter is required in the system, use a low valued series resistor ($\leq$ 1 kΩ) for a passive RC section or add an op amp RC active low pass filter. For low source resistance applications, ($\leq$ 1 kΩ), a 0.1 μF bypass capacitor at the inputs will prevent noise pickup due to series lead inductance of a long wire. A 100Ω series resistor can be used to isolate this capacitor—both the R and C are placed outside the feedback loop—from the output of an op amp, if used.

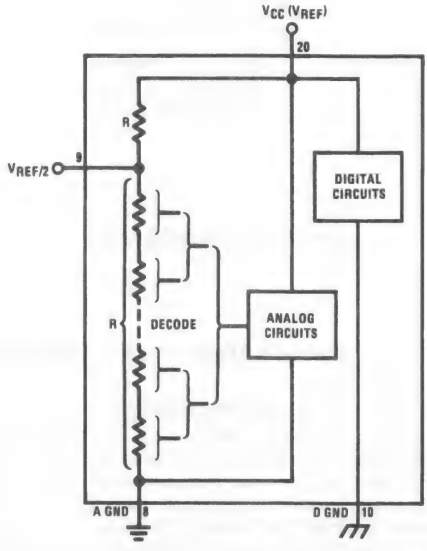
2.3.4 Noise

The leads to the analog inputs (pin 6 and 7) should be kept as short as possible to minimize input noise coupling. Both noise and undesired digital clock coupling to these inputs can cause system errors. The source resistance for these inputs should, in general, be kept below 5 kΩ. Larger values of source resistance can cause undesired system noise pickup. Input bypass capacitors, placed from the analog inputs to ground, will eliminate system noise pickup but can create analog scale errors as these capacitors will average the transient input switching currents of the A/D (see section 2.3.1.). This scale error depends on both a large source resistance and the use of an input bypass capacitor. This error can be eliminated by doing a full-scale adjustment of the A/D (adjust $V_{REF}/2$ for a proper full-scale reading—see section 2.5.2 on Full-Scale Adjustment) with the source resistance and input bypass capacitor in place.

2.4 Reference Voltage

2.4.1 Span Adjust

For maximum applications flexibility, these A/Ds have been designed to accommodate a 5 V_{DC}, 2.5 V_{DC} or an adjusted voltage reference. This has been achieved in the design of the IC as shown in *Figure 4*.



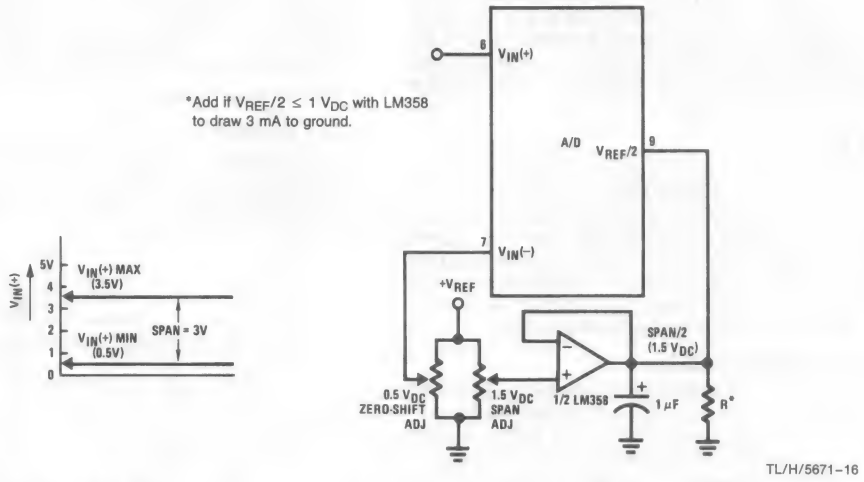
TL/H/5671–15

FIGURE 4. The $V_{REFERENCE}$ Design on the IC

Notice that the reference voltage for the IC is either ½ of the voltage applied to the V_{CC} supply pin, or is equal to the voltage that is externally forced at the $V_{REF}/2$ pin. This allows for a ratiometric voltage reference using the V_{CC} supply, a 5 V_{DC} reference voltage can be used for the V_{CC} supply or a voltage less than 2.5 V_{DC} can be applied to the $V_{REF}/2$ input for increased application flexibility. The internal gain to the $V_{REF}/2$ input is 2, making the full-scale differential input voltage twice the voltage at pin 9.

An example of the use of an adjusted reference voltage is to accommodate a reduced span—or dynamic voltage range of the analog input voltage. If the analog input voltage were to range from 0.5 V_{DC} to 3.5 V_{DC}, instead of 0V to 5 V_{DC}, the span would be 3V as shown in *Figure 5*. With 0.5 V_{DC} applied to the $V_{IN}(-)$ pin to absorb the offset, the reference voltage can be made equal to ½ of the 3V span or 1.5 V_{DC}. The A/D now will encode the $V_{IN}(+)$ signal from 0.5V to 3.5 V with the 0.5V input corresponding to zero and the 3.5 V_{DC} input corresponding to full-scale. The full 8 bits of resolution are therefore applied over this reduced analog input voltage range.

Functional Description (Continued)



a) Analog Input Signal Example

b) Accommodating an Analog Input from 0.5V (Digital Out = = 00_{HEX}) to 3.5V (Digital Out = FF_{HEX})

FIGURE 5. Adapting the A/D Analog Input Voltages to Match an Arbitrary Input Signal Range

2.4.2 Reference Accuracy Requirements

The converter can be operated in a ratiometric mode or an absolute mode. In ratiometric converter applications, the magnitude of the reference voltage is a factor in both the output of the source transducer and the output of the A/D converter and therefore cancels out in the final digital output code. The ADC0805 is specified particularly for use in ratiometric applications with no adjustments required. In absolute conversion applications, both the initial value and the temperature stability of the reference voltage are important factors in the accuracy of the A/D converter. For $V_{REF}/2$ voltages of 2.4 V_{DC} nominal value, initial errors of ±10 mV_{DC} will cause conversion errors of ±1 LSB due to the gain of 2 of the $V_{REF}/2$ input. In reduced span applications, the initial value and the stability of the $V_{REF}/2$ input voltage become even more important. For example, if the span is reduced to 2.5V, the analog input LSB voltage value is correspondingly reduced from 20 mV (5V span) to 10 mV and 1 LSB at the $V_{REF}/2$ input becomes 5 mV. As can be seen, this reduces the allowed initial tolerance of the reference voltage and requires correspondingly less absolute change with temperature variations. Note that spans smaller than 2.5V place even tighter requirements on the initial accuracy and stability of the reference source.

In general, the magnitude of the reference voltage will require an initial adjustment. Errors due to an improper value of reference voltage appear as full-scale errors in the A/D transfer function. IC voltage regulators may be used for references if the ambient temperature changes are not excessive. The LM336B 2.5V IC reference diode (from National Semiconductor) has a temperature stability of 1.8 mV typ (6 mV max) over $0°C \leq T_A \leq +70°C$. Other temperature range parts are also available.

2.5 Errors and Reference Voltage Adjustments

2.5.1 Zero Error

The zero of the A/D does not require adjustment. If the minimum analog input voltage value, $V_{IN(MIN)}$, is not ground, a zero offset can be done. The converter can be made to output 0000 0000 digital code for this minimum input voltage by biasing the A/D $V_{IN}(-)$ input at this $V_{IN(MIN)}$ value (see Applications section). This utilizes the differential mode operation of the A/D.

The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the V_{IN} (−) input and applying a small magnitude positive voltage to the V_{IN} (+) input. Zero error is the difference between the actual DC input voltage that is necessary to just cause an output digital code transition from 0000 0000 to 0000 0001 and the ideal ½ LSB value (½ LSB = 9.8 mV for $V_{REF}/2 = 2.500\ V_{DC}$).

2.5.2 Full-Scale

The full-scale adjustment can be made by applying a differential input voltage that is 1½ LSB less than the desired analog full-scale voltage range and then adjusting the magnitude of the $V_{REF}/2$ input (pin 9 or the V_{CC} supply if pin 9 is not used) for a digital output code that is just changing from 1111 1110 to 1111 1111.

Functional Description (Continued)

2.5.3 Adjusting for an Arbitrary Analog Input Voltage Range

If the analog zero voltage of the A/D is shifted away from ground (for example, to accommodate an analog input signal that does not go to ground) this new zero reference should be properly adjusted first. A $V_{IN}(+)$ voltage that equals this desired zero reference plus ½ LSB (where the LSB is calculated for the desired analog span, 1 LSB = analog span/256) is applied to pin 6 and the zero reference voltage at pin 7 should then be adjusted to just obtain the 00_{HEX} to 01_{HEX} code transition.

The full-scale adjustment should then be made (with the proper $V_{IN}(-)$ voltage applied) by forcing a voltage to the $V_{IN}(+)$ input which is given by:

$$V_{IN}(+) \text{ fs adj} = V_{MAX} - 1.5\left[\frac{(V_{MAX} - V_{MIN})}{256}\right],$$

where:

V_{MAX} = The high end of the analog input range

and

V_{MIN} = the low end (the offset zero) of the analog range. (Both are ground referenced.)

The $V_{REF}/2$ (or V_{CC}) voltage is then adjusted to provide a code change from FE_{HEX} to FF_{HEX}. This completes the adjustment procedure.

2.6 Clocking Option

The clock for the A/D can be derived from the CPU clock or an external RC can be added to provide self-clocking. The CLK IN (pin 4) makes use of a Schmitt trigger as shown in *Figure 6*.

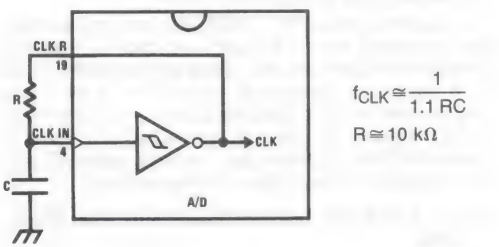


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FIGURE 6. Self-Clocking the A/D

Heavy capacitive or DC loading of the clock R pin should be avoided as this will disturb normal converter operation. Loads less than 50 pF, such as driving up to 7 A/D converter clock inputs from a single clock R pin of 1 converter, are allowed. For larger clock line loading, a CMOS or low power TTL buffer or PNP input logic should be used to minimize the loading on the clock R pin (do not use a standard TTL buffer).

2.7 Restart During a Conversion

If the A/D is restarted ($\overline{CS}$ and $\overline{WR}$ go low and return high) during a conversion, the converter is reset and a new conversion is started. The output data latch is not updated if the conversion in process is not allowed to be completed, therefore the data of the previous conversion remains in this latch. The $\overline{INTR}$ output simply remains at the "1" level.

2.8 Continuous Conversions

For operation in the free-running mode an initializing pulse should be used, following power-up, to ensure circuit operation. In this application, the $\overline{CS}$ input is grounded and the $\overline{WR}$ input is tied to the $\overline{INTR}$ output. This $\overline{WR}$ and $\overline{INTR}$ node should be momentarily forced to logic low following a power-up cycle to guarantee operation.

2.9 Driving the Data Bus

This MOS A/D, like MOS microprocessors and memories, will require a bus driver when the total capacitance of the data bus gets large. Other circuitry, which is tied to the data bus, will add to the total capacitive loading, even in TRI-STATE (high impedance mode). Backplane bussing also greatly adds to the stray capacitance of the data bus.

There are some alternatives available to the designer to handle this problem. Basically, the capacitive loading of the data bus slows down the response time, even though DC specifications are still met. For systems operating with a relatively slow CPU clock frequency, more time is available in which to establish proper logic levels on the bus and therefore higher capacitive loads can be driven (see typical characteristics curves).

At higher CPU clock frequencies time can be extended for I/O reads (and/or writes) by inserting wait states (8080) or using clock extending circuits (6800).

Finally, if time is short and capacitive loading is high, external bus drivers must be used. These can be TRI-STATE buffers (low power Schottky such as the DM74LS240 series is recommended) or special higher drive current products which are designed as bus drivers. High current bipolar bus drivers with PNP inputs are recommended.

2.10 Power Supplies

Noise spikes on the V_{CC} supply line can cause conversion errors as the comparator will respond to this noise. A low inductance tantalum filter capacitor should be used close to the converter V_{CC} pin and values of 1 μF or greater are recommended. If an unregulated voltage is available in the system, a separate LM340LAZ-5.0, TO-92, 5V voltage regulator for the converter (and other analog circuitry) will greatly reduce digital noise on the V_{CC} supply.

2.11 Wiring and Hook-Up Precautions

Standard digital wire wrap sockets are not satisfactory for breadboarding this A/D converter. Sockets on PC boards can be used and all logic signal wires and leads should be grouped and kept as far away as possible from the analog signal leads. Exposed leads to the analog inputs can cause undesired digital noise and hum pickup, therefore shielded leads may be necessary in many applications.

Functional Description (Continued)

A single point analog ground that is separate from the logic ground points should be used. The power supply bypass capacitor and the self-clocking capacitor (if used) should both be returned to digital ground. Any $V_{REF}/2$ bypass capacitors, analog input filter capacitors, or input signal shielding should be returned to the analog ground point. A test for proper grounding is to measure the zero error of the A/D converter. Zero errors in excess of ¼ LSB can usually be traced to improper board layout and wiring (see section 2.5.1 for measuring the zero error).

3.0 TESTING THE A/D CONVERTER

There are many degrees of complexity associated with testing an A/D converter. One of the simplest tests is to apply a known analog input voltage to the converter and use LEDs to display the resulting digital output code as shown in *Figure 7*.

For ease of testing, the $V_{REF}/2$ (pin 9) should be supplied with 2.560 V_{DC} and a V_{CC} supply voltage of 5.12 V_{DC} should be used. This provides an LSB value of 20 mV.

If a full-scale adjustment is to be made, an analog input voltage of 5.090 V_{DC} (5.120−1½ LSB) should be applied to the $V_{IN}(+)$ pin with the $V_{IN}(-)$ pin grounded. The value of the $V_{REF}/2$ input voltage should then be adjusted until the digital output code is just changing from 1111 1110 to 1111 1111. This value of $V_{REF}/2$ should then be used for all the tests.

The digital output LED display can be decoded by dividing the 8 bits into 2 hex characters, the 4 most significant (MS) and the 4 least significant (LS). Table I shows the fractional binary equivalent of these two 4-bit groups. By adding the voltages obtained from the "VMS" and "VLS" columns in Table I, the nominal value of the digital display (when $V_{REF}/2$ = 2.560V) can be determined. For example, for an output LED display of 1011 0110 or B6 (in hex), the voltage values from the table are 3.520 + 0.120 or 3.640 V_{DC}. These voltage values represent the center-values of a perfect A/D converter. The effects of quantization error have to be accounted for in the interpretation of the test results.

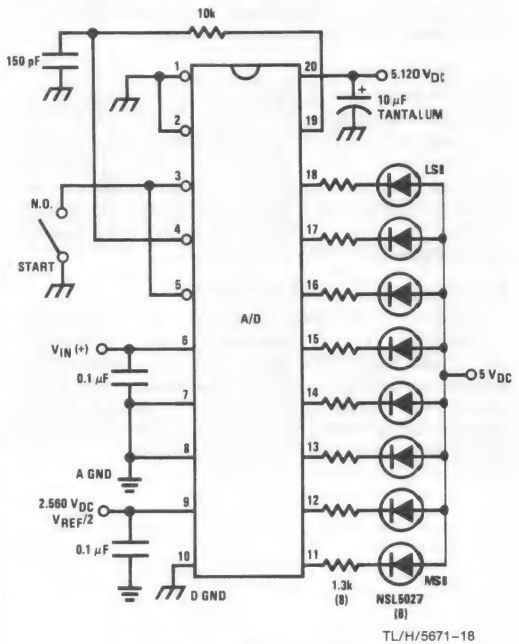

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FIGURE 7. Basic A/D Tester

For a higher speed test system, or to obtain plotted data, a digital-to-analog converter is needed for the test set-up. An accurate 10-bit DAC can serve as the precision voltage source for the A/D. Errors of the A/D under test can be expressed as either analog voltages or differences in 2 digital words.

A basic A/D tester that uses a DAC and provides the error as an analog output voltage is shown in *Figure 8*. The 2 op amps can be eliminated if a lab DVM with a numerical subtraction feature is available to read the difference voltage, "A−C", directly. The analog input voltage can be supplied by a low frequency ramp generator and an X-Y plotter can be used to provide analog error (Y axis) versus analog input (X axis).

For operation with a microprocessor or a computer-based test system, it is more convenient to present the errors digitally. This can be done with the circuit of *Figure 9*, where the output code transitions can be detected as the 10-bit DAC is incremented. This provides ¼ LSB steps for the 8-bit A/D under test. If the results of this test are automatically plotted with the analog input on the X axis and the error (in LSB's) as the Y axis, a useful transfer function of the A/D under test results. For acceptance testing, the plot is not necessary and the testing speed can be increased by establishing internal limits on the allowed error for each code.

4.0 MICROPROCESSOR INTERFACING

To dicuss the interface with 8080A and 6800 microprocessors, a common sample subroutine structure is used. The microprocessor starts the A/D, reads and stores the results of 16 successive conversions, then returns to the user's program. The 16 data bytes are stored in 16 successive memory locations. All Data and Addresses will be given in hexadecimal form. Software and hardware details are provided separately for each type of microprocessor.

4.1 Interfacing 8080 Microprocessor Derivatives (8048, 8085)

This converter has been designed to directly interface with derivatives of the 8080 microprocessor. The A/D can be mapped into memory space (using standard memory address decoding for $\overline{CS}$ and the $\overline{MEMR}$ and $\overline{MEMW}$ strobes) or it can be controlled as an I/O device by using the $\overline{I/O\ R}$ and $\overline{I/O\ W}$ strobes and decoding the address bits A0 → A7 (or address bits A8 → A15 as they will contain the same 8-bit address information) to obtain the $\overline{CS}$ input. Using the I/O space provides 256 additional addresses and may allow a simpler 8-bit address decoder but the data can only be input to the accumulator. To make use of the additional memory reference instructions, the A/D should be mapped into memory space. An example of an A/D in I/O space is shown in *Figure 10*.

Functional Description (Continued)

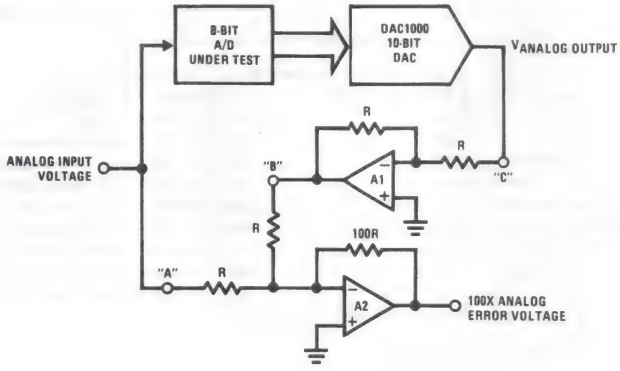

FIGURE 8. A/D Tester with Analog Error Output

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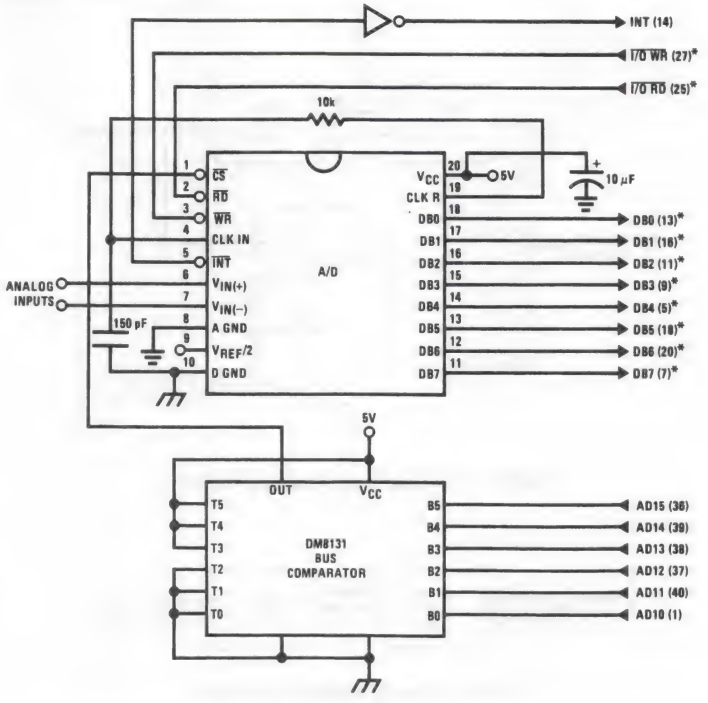
FIGURE 9. Basic "Digital" A/D Tester

TABLE I. DECODING THE DIGITAL OUTPUT LEDs

HEX	BINARY	FRACTIONAL BINARY VALUE FOR		OUTPUT VOLTAGE CENTER VALUES WITH $V_{REF}/2 = 2.560\ V_{DC}$	
		MS GROUP	LS GROUP	VMS GROUP*	VLS GROUP*
F	1 1 1 1	15/16	15/256	4.800	0.300
E	1 1 1 0	7/8	7/128	4.480	0.280
D	1 1 0 1	13/16	13/256	4.160	0.260
C	1 1 0 0	3/4	3/64	3.840	0.240
B	1 0 1 1	11/16	11/256	3.520	0.220
A	1 0 1 0	5/8	5/128	3.200	0.200
9	1 0 0 1	9/16	9/256	2/880	0.180
8	1 0 0 0	1/2	1/32	2/560	0.160
7	0 1 1 1	7/16	7/256	2.240	0.140
6	0 1 1 0	3/8	3/128	1.920	0.120
5	0 1 0 1	5/16	2/256	1.600	0.100
4	0 1 0 0	1/4	1/64	1/280	0.080
3	0 0 1 1	3/16	3/256	0.960	0.060
2	0 0 1 0	1/8	1/128	0.640	0.040
1	0 0 0 1	1/16	1/256	0.320	0.020
0	0 0 0 0			0	0

*Display Output = VMS Group + VLS Group

Functional Description (Continued)



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Note 1: *Pin numbers for the DP8228 system controller, others are INS8080A.

Note 2: Pin 23 of the INS8228 must be tied to +12V through a 1 kΩ resistor to generate the RST 7 instruction when an interrupt is acknowledged as required by the accompanying sample program.

FIGURE 10. ADC0801–INS8080A CPU Interface

SAMPLE PROGRAM FOR *FIGURE 10* ADC0801–INS8080A CPU INTERFACE

```
0038    C3 00 03    RST 7:          JMP     LD DATA
  •       •           •
  •       •           •
0100    21 00 02    START:          LXI H 0200H         ; HL pair will point to
                                                        ; data storage locations
0103    31 00 04    RETURN:         LXI SP 0400H        ; Initialize stack pointer (Note 1)
0106    7D                          MOV A, L            ; Test # of bytes entered
0107    FE 0F                       CPI 0F H            ; If # = 16. JMP to
0109    CA 13 01                    JZ CONT             ; user program
010C    D3 E0                       OUT E0 H            ; Start A/D
010E    FB                          EI                  ; Enable interrupt
010F    00          LOOP:           NOP                 ; Loop until end of
0110    C3 0F 01                    JMP LOOP            ; conversion
0113      •         CONT:             •
  •       •           •               •
  •       •         (User program to  •
  •       •         process data)     •
  •       •           •               •
  •       •           •               •
0300    DB E0       LD DATA:        IN E0 H             ; Load data into accumulator
0302    77                          MOV M, A            ; Store data
0303    23                          INX H               ; Increment storage pointer
0304    C3 03 01                    JMP RETURN
```

Note 1: The stack pointer must be dimensioned because a RST 7 instruction pushes the PC onto the stack.

Note 2: All address used were arbitrarily chosen.

Functional Description (Continued)

The standard control bus signals of the 8080 $\overline{CS}$, $\overline{RD}$ and $\overline{WR}$) can be directly wired to the digital control inputs of the A/D and the bus timing requirements are met to allow both starting the converter and outputting the data onto the data bus. A bus driver should be used for larger microprocessor systems where the data bus leaves the PC board and/or must drive capacitive loads larger than 100 pF.

4.1.1 Sample 8080A CPU Interfacing Circuitry and Program

The following sample program and associated hardware shown in *Figure 10* may be used to input data from the converter to the INS8080A CPU chip set (comprised of the INS8080A microprocessor, the INS8228 system controller and the INS8224 clock generator). For simplicity, the A/D is controlled as an I/O device, specifically an 8-bit bi-directional port located at an arbitrarily chosen port address, E0. The TRI-STATE output capability of the A/D eliminates the need for a peripheral interface device, however address decoding is still required to generate the appropriate $\overline{CS}$ for the converter.

It is important to note that in systems where the A/D converter is 1-of-8 or less I/O mapped devices, no address decoding circuitry is necessary. Each of the 8 address bits (A0 to A7) can be directly used as $\overline{CS}$ inputs—one for each I/O device.

4.1.2 INS8048 Interface

The INS8048 interface technique with the ADC0801 series (see *Figure 11*) is simpler than the 8080A CPU interface. There are 24 I/O lines and three test input lines in the 8048. With these extra I/O lines available, one of the I/O lines (bit 0 of port 1) is used as the chip select signal to the A/D, thus eliminating the use of an external address decoder. Bus control signals $\overline{RD}$, $\overline{WR}$ and $\overline{INT}$ of the 8048 are tied directly to the A/D. The 16 converted data words are stored at on-chip RAM locations from 20 to 2F (Hex). The $\overline{RD}$ and $\overline{WR}$ signals are generated by reading from and writing into a dummy address, respectively. A sample interface program is shown below.

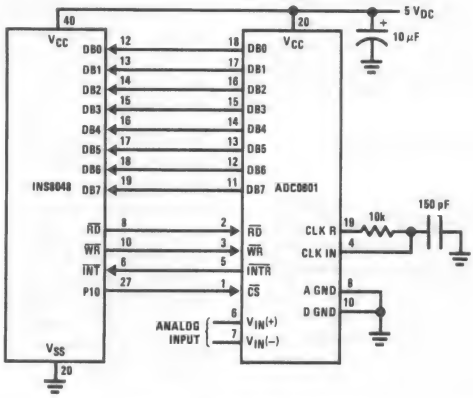

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FIGURE 11. INS8048 Interface

SAMPLE PROGRAM FOR *FIGURE 11* INS8048 INTERFACE

```
04 10                   JMP     10H             ; Program starts at addr 10
                        ORG     3H
04 50                   JMP     50H             ; Interrupt jump vector
                        ORG     10H             ; Main program
99 FE                   ANL     P1, #0FEH       ; Chip select
81                      MOVX    A, @R1          ; Read in the 1st data
                                                ; to reset the intr
89 01       START:      ORL     P1, #1          ; Set port pin high
B8 20                   MOV     R0, #20H        ; Data address
B9 FF                   MOV     R1, #0FFH       ; Dummy address
BA 10                   MOV     R2, #10H        ; Counter for 16 bytes
23 FF       AGAIN:      MOV     A, #0FFH        ; Set ACC for intr loop
99 FE                   ANL     P1, #0FEH       ; Send CS (bit 0 of P1)
91                      MOVX    @R1, A          ; Send WR out
05                      EN      I               ; Enable interrupt
96 21       LOOP:       JNZ     LOOP            ; Wait for interrupt
EA 1B                   DJNZ    R2, AGAIN       ; If 16 bytes are read
00                      NOP                     ; go to user's program
00                      NOP
                        ORG     50H
81          INDATA:     MOVX    A, @R1          ; Input data, CS still low
A0                      MOV     @R0, A          ; Store in memory
18                      INC     R0              ; Increment storage counter
89 01                   ORL     P1, #1          ; Reset CS signal
27                      CLR     A               ; Clear ACC to get out of
93                      RETR                    ; the interrupt loop
```

Functional Description (Continued)

4.2 Interfacing the Z-80

The Z-80 control bus is slightly different from that of the 8080. General $\overline{RD}$ and $\overline{WR}$ strobes are provided and separate memory request, $\overline{MREQ}$, and I/O request, $\overline{IORQ}$, signals are used which have to be combined with the generalized strobes to provide the equivalent 8080 signals. An advantage of operating the A/D in I/O space with the Z-80 is that the CPU will automatically insert one wait state (the $\overline{RD}$ and $\overline{WR}$ strobes are extended one clock period) to allow more time for the I/O devices to respond. Logic to map the A/D in I/O space is shown in *Figure 13*.

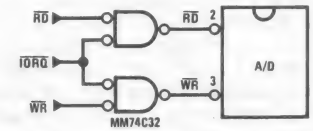

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FIGURE 13. Mapping the A/D as an I/O Device for Use with the Z-80 CPU

Additional I/O advantages exist as software DMA routines are available and use can be made of the output data transfer which exists on the upper 8 address lines (A8 to A15) during I/O input instructions. For example, MUX channel selection for the A/D can be accomplished with this operating mode.

4.3 Interfacing 6800 Microprocessor Derivatives (6502, etc.)

The control bus for the 6800 microprocessor derivatives does not use the $\overline{RD}$ and $\overline{WR}$ strobe signals. Instead it employs a single R/$\overline{W}$ line and additional timing, if needed, can be derived fom the $\phi 2$ clock. All I/O devices are memory mapped in the 6800 system, and a special signal, VMA, indicates that the current address is valid. *Figure 14* shows an interface schematic where the A/D is memory mapped in the 6800 system. For simplicity, the $\overline{CS}$ decoding is shown using ½ DM8092. Note that in many 6800 systems, an already decoded $\overline{4/5}$ line is brought out to the common bus at pin 21. This can be tied directly to the $\overline{CS}$ pin of the A/D, provided that no other devices are addressed at HX ADDR: 4XXX or 5XXX.

The following subroutine performs essentially the same function as in the case of the 8080A interface and it can be called from anywhere in the user's program.

In *Figure 15* the ADC0801 series is interfaced to the M6800 microprocessor through (the arbitrarily chosen) Port B of the MC6820 or MC6821 Peripheral Interface Adapter, (PIA). Here the $\overline{CS}$ pin of the A/D is grounded since the PIA is already memory mapped in the M6800 system and no $\overline{CS}$ decoding is necessary. Also notice that the A/D output data lines are connected to the microprocessor bus under program control through the PIA and therefore the A/D $\overline{RD}$ pin can be grounded.

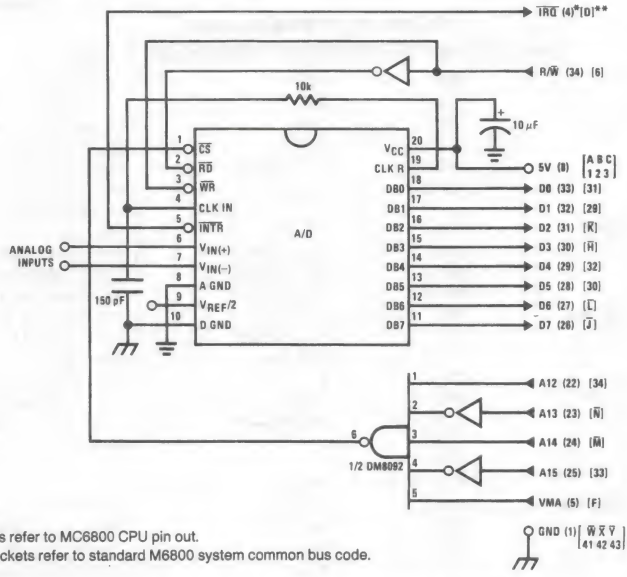
A sample interface program equivalent to the previous one is shown below *Figure 15*. The PIA Data and Control Registers of Port B are located at HEX addresses 8006 and 8007, respectively.

5.0 GENERAL APPLICATIONS

The following applications show some interesting uses for the A/D. The fact that one particular microprocessor is used is not meant to be restrictive. Each of these application circuits would have its counterpart using any microprocessor that is desired.

5.1 Multiple ADC0801 Series to MC6800 CPU Interface

To transfer analog data from several channels to a single microprocessor system, a multiple converter scheme presents several advantages over the conventional multiplexer single-converter approach. With the ADC0801 series, the differential inputs allow individual span adjustment for each channel. Furthermore, all analog input channels are sensed simultaneously, which essentially divides the microprocessor's total system servicing time by the number of channels, since all conversions occur simultaneously. This scheme is shown in *Figure 16*.



Note 1: Numbers in parentheses refer to MC6800 CPU pin out.
Note 2: Number or letters in brackets refer to standard M6800 system common bus code.

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FIGURE 14. ADC0801-MC6800 CPU Interface

Functional Description (Continued)

SAMPLE PROGRAM FOR *FIGURE 14* ADC0801-MC6800 CPU INTERFACE

```
0010  DF 36      DATAIN  STX   TEMP2   ; Save contents of X
0012  CE 00 2C            LDX   #$002C  ; Upon IRQ low CPU
0015  FF FF F8            STX   $FFF8   ; jumps to 002C
0018  B7 50 00            STAA  $5000   ; Start ADC0801
001B  0E                  CLI
001C  3E          CONVRT  WAI           ; Wait for interrupt
001D  DE 34               LDX   TEMP1
001F  8C 02 0F            CPX   #$020F  ; Is final data stored?
0022  27 14               BEQ   ENDP
0024  B7 50 00            STAA  $5000   ; Restarts ADC0801
0027  08                  INX
0028  DF 34               STX   TEMP1
002A  20 F0               BRA   CONVRT
002C  DE 34       INTRPT  LDX   TEMP1
002E  B6 50 00            LDAA  $5000   ; Read data
0031  A7 00               STAA  X       ; Store it at X
0033  3B                  RTI
0034  02 00       TEMP1   FDB   $0200   ; Starting address for
                                        ; data storage
0036  00 00       TEMP2   FDB   $0000
0038  CE 02 00    ENDP    LDX   #$0200  ; Reinitialize TEMP1
003B  DF 34               STX   TEMP1
003D  DE 36               LDX   TEMP2
003F  39                  RTS           ; Return from subroutine
                                        ; To user's program
```

Note 1: In order for the microprocessor to service subroutines and interrupts, the stack pointer must be dimensioned in the user's program.

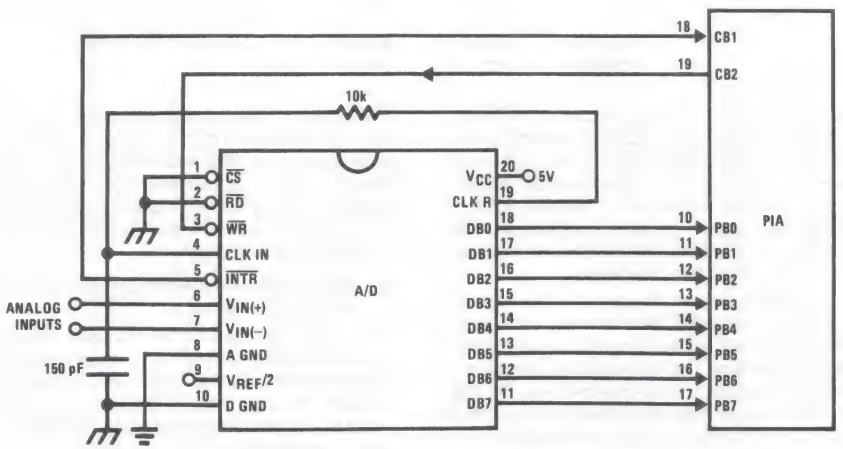

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FIGURE 15. ADC0801–MC6820 PIA Interface

Functional Description (Continued)

SAMPLE PROGRAM FOR *FIGURE 15* ADC0801–MC6820 PIA INTERFACE

```
0010    CE 00 38    DATAIN    LDX    #$0038    ; Upon IRQ low CPU
0013    FF FF F8              STX    $FFF8     ; jumps to 0038
0016    B6 80 06              LDAA   PIAORB    ; Clear possible IRQ flags
0019    4F                    CLRA
001A    B7 80 07              STAA   PIACRB
001D    B7 80 06              STAA   PIAORB    ; Set Port B as input
0020    0E                    CLI
0021    C6 34                 LDAB   #$34
0023    86 3D                 LDAA   #$3D
0025    F7 80 07    CONVRT    STAB   PIACRB    ; Starts ADC0801
0028    B7 80 07              STAA   PIACRB
002B    3E                    WAI              ; Wait for interrupt
002C    DE 40                 LDX    TEMP1
002E    8C 02 0F              CPX    #$020F    ; Is final data stored?
0031    27 0F                 BEQ    ENDP
0033    08                    INX
0034    DF 40                 STX    TEMP1
0036    20 ED                 BRA    CONVRT
0038    DE 40       INTRPT    LDX    TEMP1
003A    B6 80 06              LDAA   PIAORB    ; Read data in
003D    A7 00                 STAA   X         ; Store it at X
003F    3B                    RTI
0040    02 00       TEMP1     FDB    $0200     ; Starting address for
                                                ; data storage
0042    CE 02 00    ENDP      LDX    #$0200    ; Reinitialize TEMP1
0045    DF 40                 STX    TEMP1
0047    39                    RTS              ; Return from subroutine
                    PIAORB    EQU    $8006     ; To user's program
                    PIACRB    EQU    $8007
```

The following schematic and sample subroutine (DATA IN) may be used to interface (up to) 8 ADC0801's directly to the MC6800 CPU. This scheme can easily be extended to allow the interface of more converters. In this configuration the converters are (arbitrarily) located at HEX address 5000 in the MC6800 memory space. To save components, the clock signal is derived from just one RC pair on the first converter. This output drives the other A/Ds.

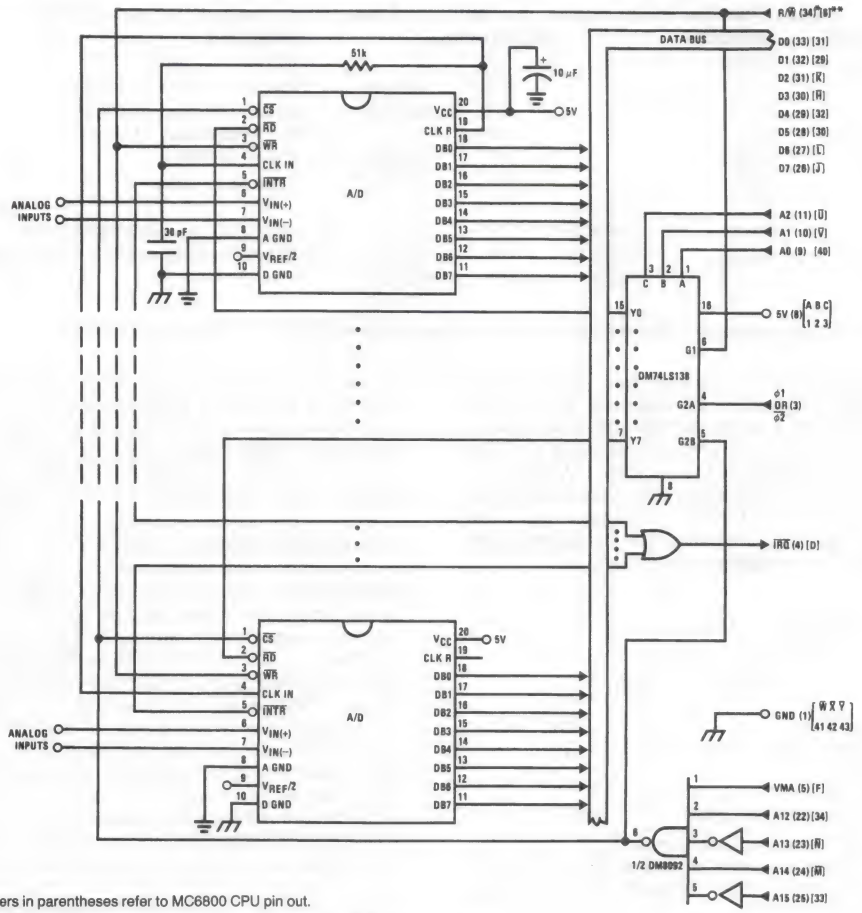
All the converters are started simultaneously with a STORE instruction at HEX address 5000. Note that any other HEX address of the form 5XXX will be decoded by the circuit, pulling all the $\overline{CS}$ inputs low. This can easily be avoided by using a more definitive address decoding scheme. All the interrupts are ORed together to insure that all A/Ds have completed their conversion before the microprocessor is interrupted.

The subroutine, DATA IN, may be called from anywhere in the user's program. Once called, this routine initializes the CPU, starts all the converters simultaneously and waits for the interrupt signal. Upon receiving the interrupt, it reads the converters (from HEX addresses 5000 through 5007) and stores the data successively at (arbitrarily chosen) HEX addresses 0200 to 0207, before returning to the user's program. All CPU registers then recover the original data they had before servicing DATA IN.

5.2 Auto-Zeroed Differential Transducer Amplifier and A/D Converter

The differential inputs of the ADC0801 series eliminate the need to perform a differential to single ended conversion for a differential transducer. Thus, one op amp can be eliminated since the differential to single ended conversion is provided by the differential input of the ADC0801 series. In general, a transducer preamp is required to take advantage of the full A/D converter input dynamic range.

Functional Description (Continued)



Note 1: Numbers in parentheses refer to MC6800 CPU pin out.

Note 2: Numbers of letters in brackets refer to standard M6800 system common bus code.

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FIGURE 16. Interfacing Multiple A/Ds in an MC6800 System

SAMPLE PROGRAM FOR *FIGURE 16* INTERFACING MULTIPLE A/Ds IN AN MC6800 SYSTEM

ADDRESS	HEX CODE		MNEMONICS		COMMENTS
0010	DF 44	DATAIN	STX	TEMP	; Save Contents of X
0012	CE 00 2A		LDX	#$002A	; Upon $\overline{IRQ}$ LOW CPU
0015	FF FF F8		STX	$FFF8	; Jumps to 002A
0018	B7 50 00		STAA	$5000	; Starts all A/D's
001B	0E		CLI		
001C	3E		WAI		; Wait for interrupt
001D	CE 50 00		LDX	#$5000	
0020	DF 40		STX	INDEX1	; Reset both INDEX
0022	CE 02 00		LDX	#$0200	; 1 and 2 to starting
0025	DF 42		STX	INDEX2	; addresses
0027	DE 44		LDX	TEMP	
0029	39		RTS		; Return from subroutine
002A	DE 40	INTRPT	LDX	INDEX1	; INDEX1 → X
002C	A6 00		LDAA	X	; Read data in from A/D at X
002E	08		INX		; Increment X by one
002F	DF 40		STX	INDEX1	; X → INDEX1
0031	DE 42		LDX	INDEX2	; INDEX2 → X

Functional Description (Continued)

SAMPLE PROGRAM FOR *FIGURE 16* INTERFACING MULTIPLE A/Ds IN AN MC6800 SYSTEM

ADDRESS	HEX CODE		MNEMONICS		COMMENTS
0033	A7 00		STAA	X	; Store data at X
0035	8C 02 07		CPX	#$0207	; Have all A/D's been read?
0038	27 05		BEQ	RETURN	; Yes: branch to RETURN
003A	08		INX		; No: increment X by one
003B	DF 42		STX	INDEX2	; X → INDEX2
003D	20 EB		BRA	INTRPT	; Branch to 002A
003F	3B	RETURN	RTI		
0040	50 00	INDEX1	FDB	$5000	; Starting address for A/D
0042	02 00	INDEX2	FDB	$0200	; Starting address for data storage
0044	00 00	TEMP	FDB	$0000	

Note 1: In order for the microprocessor to service subroutines and interrupts, the stack pointer must be dimensioned in the user's program.

For amplification of DC input signals, a major system error is the input offset voltage of the amplifiers used for the preamp. *Figure 17* is a gain of 100 differential preamp whose offset voltage errors will be cancelled by a zeroing subroutine which is performed by the INS8080A microprocessor system. The total allowable input offset voltage error for this preamp is only 50 μV for ¼ LSB error. This would obviously require very precise amplifiers. The expression for the differential output voltage of the preamp is:

$$V_O = \underbrace{[V_{IN}(+) - V_{IN}(-)]}_{\text{SIGNAL}} \underbrace{\left[1 + \frac{2R2}{R1}\right]}_{\text{GAIN}} +$$

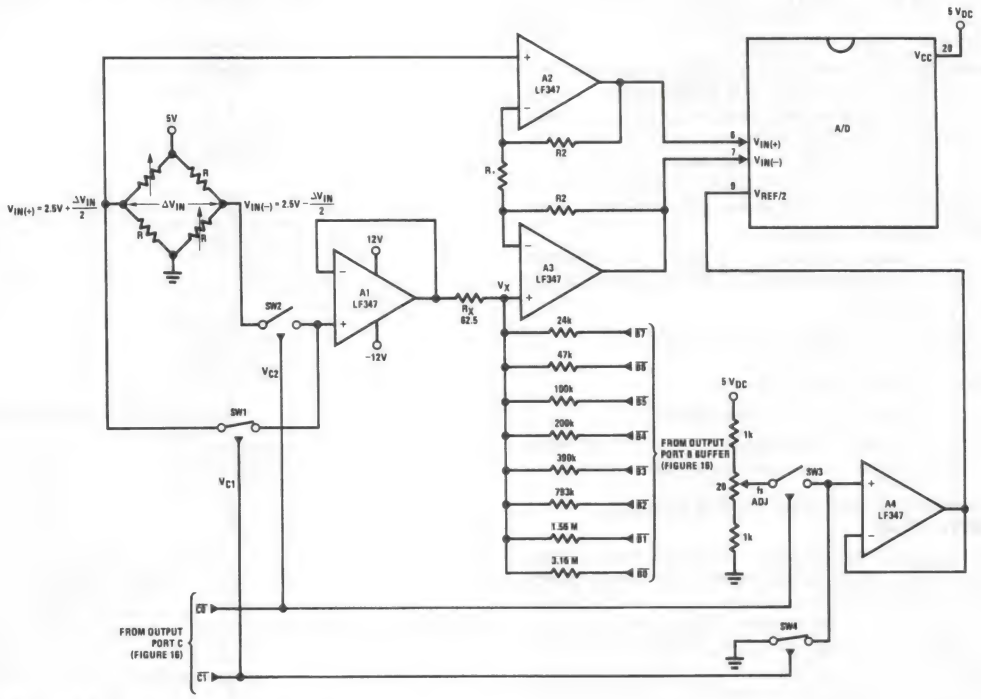
$$\underbrace{(V_{OS_2} - V_{OS_1} - V_{OS_3} \pm I_X R_X)}_{\text{DC ERROR TERM}} \underbrace{\left(1 + \frac{2R2}{R1}\right)}_{\text{GAIN}}$$

where I_X is the current through resistor R_X. All of the offset error terms can be cancelled by making $\pm I_X R_X = V_{OS1} + V_{OS3} - V_{OS2}$. This is the principle of this auto-zeroing scheme.

The INS8080A uses the 3 I/O ports of an INS8255 Programable Peripheral Interface (PPI) to control the auto zeroing and input data from the ADC0801 as shown in *Figure 18*. The PPI is programmed for basic I/O operation (mode 0) with Port A being an input port and Ports B and C being output ports. Two bits of Port C are used to alternately open or close the 2 switches at the input of the preamp. Switch SW1 is closed to force the preamp's differential input to be zero during the zeroing subroutine and then opened and SW2 is then closed for conversion of the actual differential input signal. Using 2 switches in this manner eliminates concern for the ON resistance of the switches as they must conduct only the input bias current of the input amplifiers.

Output Port B is used as a successive approximation register by the 8080 and the binary scaled resistors in series with each output bit create a D/A converter. During the zeroing subroutine, the voltage at V_X increases or decreases as required to make the differential output voltage equal to zero. This is accomplished by ensuring that the voltage at the output of A1 is approximately 2.5V so that a logic "1" (5V) on any output of Port B will source current into node V_X thus raising the voltage at V_X and making the output differential more negative. Conversely, a logic "0" (0V) will pull current out of node V_X and decrease the voltage, causing the differential output to become more positive. For the resistor values shown, V_X can move $\pm$12 mV with a resolution of 50 μV, which will null the offset error term to ¼ LSB of full-scale for the ADC0801. It is important that the voltage levels that drive the auto-zero resistors be constant. Also, for symmetry, a logic swing of 0V to 5V is convenient. To achieve this, a CMOS buffer is used for the logic output signals of Port B and this CMOS package is powered with a stable 5V source. Buffer amplifier A1 is necessary so that it can source or sink the D/A output current.

Functional Description (Continued)



Note 1: R2 = 49.5 R1

Note 2: Switches are LMC13334 CMOS analog switches.

Note 3: The 9 resistors used in the auto-zero section can be ±5% tolerance.

FIGURE 17. Gain of 100 Differential Transducer Preamp

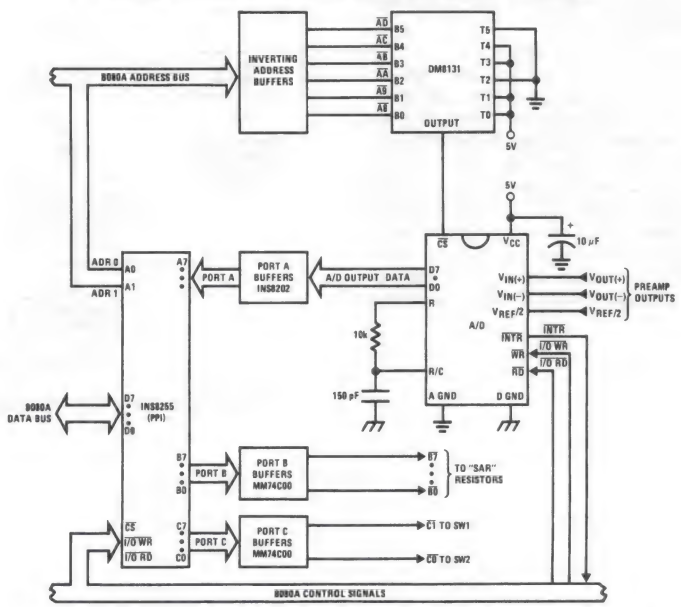


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FIGURE 18. Microprocessor Interface Circuitry for Differential Preamp

A flow chart for the zeroing subroutine is shown in *Figure 19*. It must be noted that the ADC0801 series will output an all zero code when it converts a negative input [$V_{IN}(-) \geq V_{IN}(+)$]. Also, a logic inversion exists as all of the I/O ports are buffered with inverting gates.

Basically, if the data read is zero, the differential output voltage is negative, so a bit in Port B is cleared to pull V_X more negative which will make the output more positive for the next conversion. If the data read is not zero, the output voltage is positive so a bit in Port B is set to make V_X more positive and the output more negative. This continues for 8 approximations and the differential output eventually converges to within 5 mV of zero.

The actual program is given in *Figure 20*. All addresses used are compatible with the BLC 80/10 microcomputer system. In particular:

Port A and the ADC0801 are at port address E4

Port B is at port address E5

Port C is at port address E6

PPI control word port is at port address E7

Program Counter automatically goes to ADDR:3C3D upon acknowledgement of an interrupt from the ADC0801

5.3 Multiple A/D Converters in a Z-80 Interrupt Driven Mode

In data acquisition systems where more than one A/D converter (or other peripheral device) will be interrupting program execution of a microprocessor, there is obviously a need for the CPU to determine which device requires servicing. *Figure 21* and the accompanying software is a method of determining which of 7 ADC0801 converters has completed a conversion ($\overline{INTR}$ asserted) and is requesting an interrupt. This circuit allows starting the A/D converters in any sequence, but will input and store valid data from the converters with a priority sequence of A/D 1 being read first, A/D 2 second, etc., through A/D 7 which would have the lowest priority for data being read. Only the converters whose INT is asserted will be read.

The key to decoding circuitry is the DM74LS373, 8-bit D type flip-flop. When the Z-80 acknowledges the interrupt, the program is vectored to a data input Z-80 subroutine. This subroutine will read a peripheral status word from the DM74LS373 which contains the logic state of the $\overline{INTR}$ outputs of all the converters. Each converter which initiates an interrupt will place a logic "0" in a unique bit position in the status word and the subroutine will determine the identity of the converter and execute a data read. An identifier word (which indicates which A/D the data came from) is stored in the next sequential memory location above the location of the data so the program can keep track of the identity of the data entered.

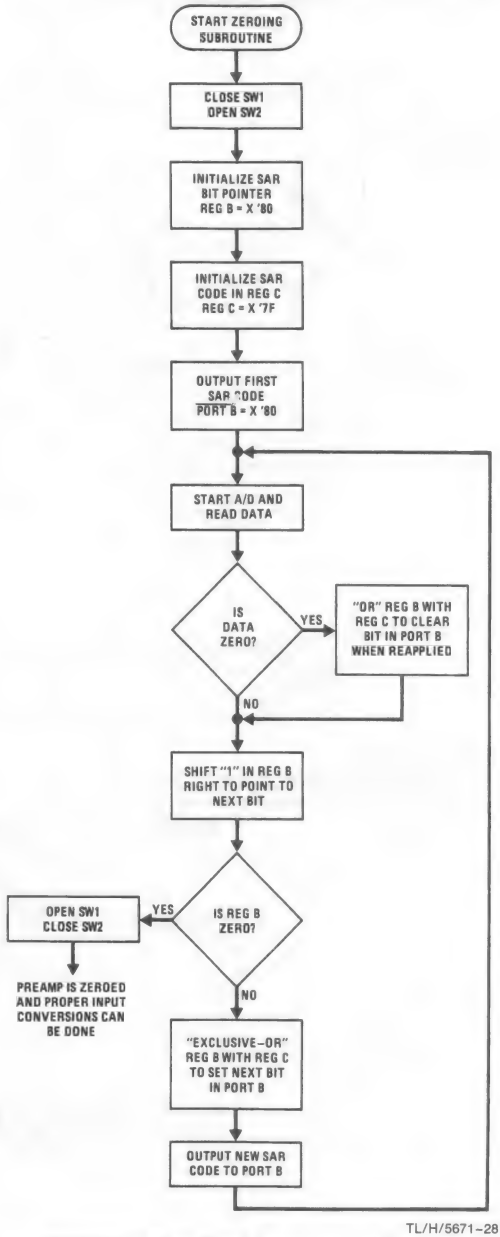
START ZEROING SUBROUTINE
CLOSE SW1 OPEN SW2
INITIALIZE SAR BIT POINTER REG B = X '80
INITIALIZE SAR CODE IN REG C REG C = X '7F
OUTPUT FIRST SAR CODE PORT B = X '80
START A/D AND READ DATA
IS DATA ZERO?
YES
"OR" REG B WITH REG C TO CLEAR BIT IN PORT B WHEN REAPPLIED
NO
SHIFT "1" IN REG B RIGHT TO POINT TO NEXT BIT
IS REG B ZERO?
YES
OPEN SW1 CLOSE SW2
PREAMP IS ZEROED AND PROPER INPUT CONVERSIONS CAN BE DONE
NO
"EXCLUSIVE-OR" REG B WITH REG C TO SET NEXT BIT IN PORT B
OUTPUT NEW SAR CODE TO PORT B

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FIGURE 19. Flow Chart for Auto-Zero Routine

```
3D00   3E90     MVI 90
3D02   D3E7     Out Control Port                          ; Program PPI
3D04   2601     MVI H 01          Auto-Zero Subroutine
3D06   7C       MOV A,H
3D07   D3E6     OUT C                                    ; Close SW1 open SW2
3D09   0680     MVI B 80                                 ; Initialize SAR bit pointer
3D0B   3E7F     MVI A 7F                                 ; Initialize SAR code
3D0D   4F       MOV C,A           Return
3D0E   D3E5     OUT B                                    ; Port B = SAR code
3D10   31AA3D   LXI SP 3DAA       Start                  ; Dimension stack pointer
3D13   D3E4     OUT A                                    ; Start A/D
3D15   FB       IE
3D16   00       NOP               Loop                   ; Loop until INT asserted
3D17   C3163D   JMP Loop
3D1A   7A       MOV A,D           Auto-Zero
3D1B   C600     ADI 00
3D1D   CA2D3D   JZ Set C                                 ; Test A/D output data for zero
3D20   78       MOV A,B           Shift B
3D21   F600     ORI 00                                   ; Clear carry
3D23   1F       RAR                                      ; Shift "1" in B right one place
3D24   FE00     CPI 00                                   ; Is B zero? If yes last
3D26   CA373D   JZ Done                                  ; approximation has been made
3D29   47       MOV B,A
3D2A   C3333D   JMP New C
3D2D   79       MOV A,C           Set C
3D2E   B0       ORA B                                    ; Set bit in C that is in same
3D2F   4F       MOV C,A                                  ; position as "1" in B
3D30   C3203D   JMP Shift B
3D33   A9       XRA C             New C                  ; Clear bit in C that is in
3D34   C30D3D   JMP Return                               ; same position as "1" in B
3D37   47       MOV B,A           Done                   ; then output new SAR code.
3D38   7C       MOV A,H                                  ; Open SW1, close SW2 then
3D39   EE03     XRI 03                                   ; proceed with program. Preamp
3D3B   D3E6     OUT C                                    ; is now zeroed.
3D3D            •                 Normal
                •
                •
                Program for processing
                proper data values
3C3D   DBE4     IN A              Read A/D Subroutine    ; Read A/D data
3C3F   EEFF     XRI FF                                   ; Invert data
3C41   57       MOV D,A
3C42   78       MOV A,B                                  ; Is B Reg = 0? If not stay
3C43   E6FF     ANI FF                                   ; in auto zero subroutine
3C45   C21A3D   JNZ Auto-Zero
3C48   C33D3D   JMP Normal
```

Note: All numerical values are hexadecimal representations.

FIGURE 20. Software for Auto-Zeroed Differential A/D

5.3 Multiple A/D Converters in a Z-80® Interrupt Driven Mode (Continued)

The following notes apply:

1) It is assumed that the CPU automatically performs a RST 7 instruction when a valid interrupt is acknowledged (CPU is in interrupt mode 1). Hence, the subroutine starting address of X0038.

2) The address bus from the Z-80 and the data bus to the Z-80 are assumed to be inverted by bus drivers.

3) A/D data and identifying words will be stored in sequential memory locations starting at the arbitrarily chosen address X 3E00.

4) The stack pointer must be dimensioned in the main program as the RST 7 instruction automatically pushes the PC onto the stack and the subroutine uses an additional 6 stack addresses.

5) The peripherals of concern are mapped into I/O space with the following port assignments:

HEX PORT ADDRESS	PERIPHERAL
00	MM74C374 8-bit flip-flop
01	A/D 1
02	A/D 2
03	A/D 3
04	A/D 4
05	A/D 5
06	A/D 6
07	A/D 7

This port address also serves as the A/D identifying word in the program.

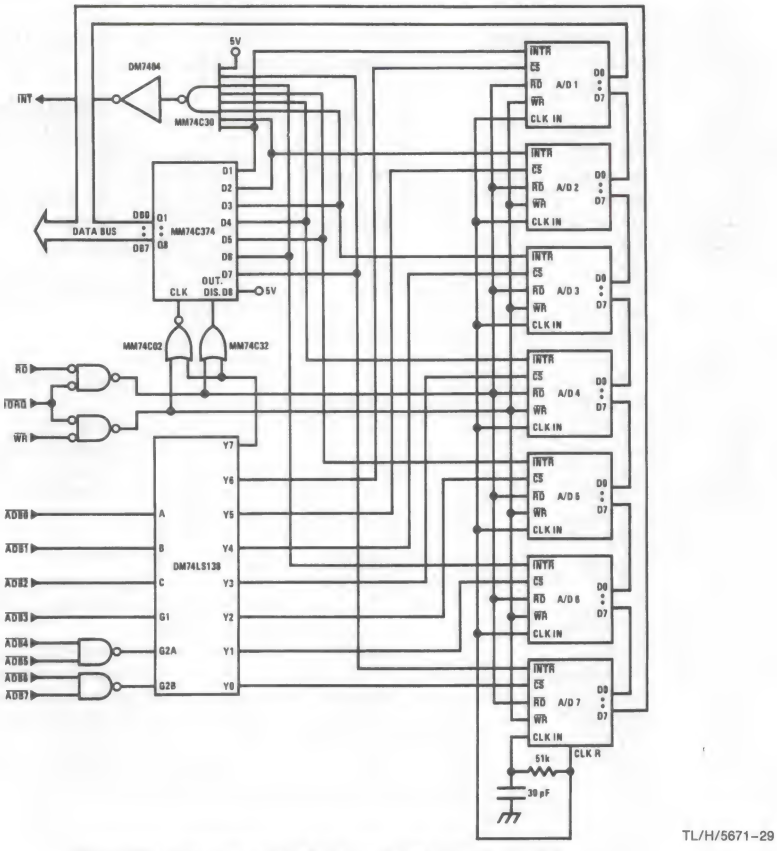

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FIGURE 21. Multiple A/Ds with Z-80 Type Microprocessor

INTERRUPT SERVICING SUBROUTINE

LOC	OBJ CODE	SOURCE STATEMENT	COMMENT
0038	E5	PUSH HL	; Save contents of all registers affected by
0039	C5	PUSH BC	; this subroutine.
003A	F5	PUSH AF	; Assumed INT mode 1 earlier set.
003B	21 00 3E	LD (HL),X3E00	; Initialize memory pointer where data will be stored.
003E	0E 01	LD C, X01	; C register will be port ADDR of A/D converters.
0040	D300	OUT X00, A	; Load peripheral status word into 8-bit latch.
0042	DB00	IN A, X00	; Load status word into accumulator.
0044	47	LD B,A	; Save the status word.
0045	79	TEST LD A,C	; Test to see if the status of all A/D's have
0046	FE 08	CP, X08	; been checked. If so, exit subroutine
0048	CA 60 00	JPZ, DONE	
004B	78	LD A,B	; Test a single bit in status word by looking for
004C	1F	RRA	; a "1" to be rotated into the CARRY (an $\overline{INT}$
004D	47	LD B,A	; is loaded as a "1"). If CARRY is set then load
004E	DA 5500	JPC, LOAD	; contents of A/D at port ADDR in C register.
0051	0C	NEXT INC C	; If CARRY is not set, increment C register to point
0052	C3 4500	JP,TEST	; to next A/D, then test next bit in status word.
0055	ED 78	LOAD IN A, (C)	; Read data from interrupting A/D and invert
0057	EE FF	XOR FF	; the data.
0059	77	LD (HL),A	; Store the data
005A	2C	INC L	
005B	71	LD (HL),C	; Store A/D identifier (A/D port ADDR).
005C	2C	INC L	
005D	C3 51 00	JP,NEXT	; Test next bit in status word.
0060	F1	DONE POP AF	; Re-establish all registers as they were
0061	C1	POP BC	; before the interrupt.
0062	E1	POP HL	
0063	C9	RET	; Return to original program

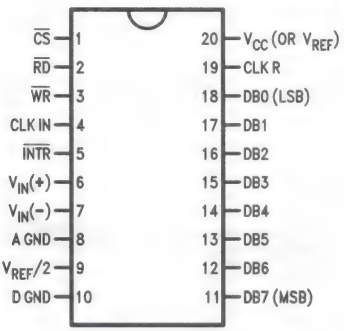
Ordering Information

TEMP RANGE		0°C TO 70°C	0°C TO 70°C	0°C TO 70°C	−40°C TO +85°C
ERROR	±¼ Bit Adjusted				ADC0801LCN
	±½ Bit Unadjusted	ADC0802LCWM	ADC0802LCV		ADC0802LCN
	±½ Bit Adjusted	ADC0803LCWM	ADC0803LCV		ADC0803LCN
	±1Bit Unadjusted	ADC0804LCWM	ADC0804LCV	ADC0804LCN	ADC0805LCN
PACKAGE OUTLINE		M20B—Small Outline	V20A—Chip Carrier	N20A—Molded DIP	

TEMP RANGE		−40°C TO +85°C	−55°C TO +125°C
ERROR	±¼ Bit Adjusted	ADC0801LCJ	ADC0801LJ
	±½ Bit Unadjusted	ADC0802LCJ	ADC0802LJ
	±½ Bit Adjusted	ADC0803LCJ	
	±1Bit Unadjusted	ADC0804LCJ	
PACKAGE OUTLINE		J20A—Cavity DIP	J20A—Cavity DIP

Connection Diagrams

ADC080X
Dual-In-Line and Small Outline (SO) Packages



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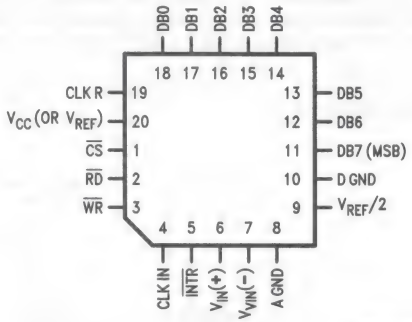
ADC080X
Molded Chip Carrier (PCC) Package



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See Ordering Information

National Semiconductor

ADC0808/ADC0809 8-Bit μP Compatible A/D Converters with 8-Channel Multiplexer

General Description

The ADC0808, ADC0809 data acquisition component is a monolithic CMOS device with an 8-bit analog-to-digital converter, 8-channel multiplexer and microprocessor compatible control logic. The 8-bit A/D converter uses successive approximation as the conversion technique. The converter features a high impedance chopper stabilized comparator, a 256R voltage divider with analog switch tree and a successive approximation register. The 8-channel multiplexer can directly access any of 8-single-ended analog signals.

The device eliminates the need for external zero and full-scale adjustments. Easy interfacing to microprocessors is provided by the latched and decoded multiplexer address inputs and latched TTL TRI-STATE® outputs.

The design of the ADC0808, ADC0809 has been optimized by incorporating the most desirable aspects of several A/D conversion techniques. The ADC0808, ADC0809 offers high speed, high accuracy, minimal temperature dependence, excellent long-term accuracy and repeatability, and consumes minimal power. These features make this device ideally suited to applications from process and machine control to consumer and automotive applications. For 16-channel multiplexer with common output (sample/hold port) see ADC0816 data sheet. (See AN-247 for more information.)

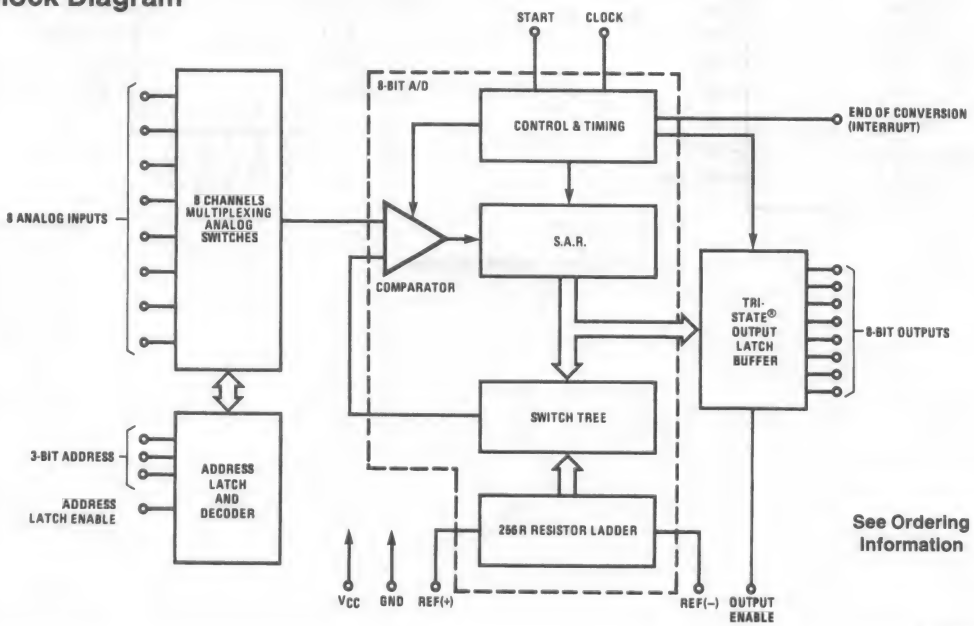
Features

- Easy interface to all microprocessors
- Operates ratiometrically or with 5 V_{DC} or analog span adjusted voltage reference
- No zero or full-scale adjust required
- 8-channel multiplexer with address logic
- 0V to 5V input range with single 5V power supply
- Outputs meet TTL voltage level specifications
- Standard hermetic or molded 28-pin DIP package
- 28-pin molded chip carrier package
- ADC0808 equivalent to MM74C949
- ADC0809 equivalent to MM74C949-1

Key Specifications

■ Resolution	8 Bits
■ Total Unadjusted Error	±½ LSB and ±1 LSB
■ Single Supply	5 V_{DC}
■ Low Power	15 mW
■ Conversion Time	100 μs

Block Diagram



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Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC}) (Note 3)	6.5V
Voltage at Any Pin Except Control Inputs	$-0.3V$ to ($V_{CC}+0.3V$)
Voltage at Control Inputs (START, OE, CLOCK, ALE, ADD A, ADD B, ADD C)	$-0.3V$ to $+15V$
Storage Temperature Range	$-65°C$ to $+150°C$
Package Dissipation at $T_A=25°C$	875 mW
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C
ESD Susceptibility (Note 11)	400V

Operating Conditions (Notes 1 & 2)

Temperature Range (Note 1)	$T_{MIN}\le T_A\le T_{MAX}$
ADC0808CJ	$-55°C\le T_A\le +125°C$
ADC0808CCJ, ADC0808CCN, ADC0809CCN	$-40°C\le T_A\le +85°C$
ADC0808CCV, ADC0809CCV	$-40°C\le T_A\le +85°C$
Range of V_{CC} (Note 1)	4.5 V_{DC} to 6.0 V_{DC}

Electrical Characteristics

Converter Specifications: $V_{CC}=5$ $V_{DC}=V_{REF+}$, $V_{REF(-)}=GND$, $T_{MIN}\le T_A\le T_{MAX}$ and $f_{CLK}=640$ kHz unless otherwise stated.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
	ADC0808					
	Total Unadjusted Error	25°C			±½	LSB
	(Note 5)	T_{MIN} to T_{MAX}			±¾	LSB
	ADC0809					
	Total Unadjusted Error	0°C to 70°C			±1	LSB
	(Note 5)	T_{MIN} to T_{MAX}			±1¼	LSB
	Input Resistance	From Ref(+) to Ref(−)	1.0	2.5		kΩ
	Analog Input Voltage Range	(Note 4) V(+) or V(−)	GND−0.10		$V_{CC}+0.10$	V_{DC}
$V_{REF(+)}$	Voltage, Top of Ladder	Measured at Ref(+)		V_{CC}	$V_{CC}+0.1$	V
$\frac{V_{REF(+)}+V_{REF(-)}}{2}$	Voltage, Center of Ladder		$V_{CC}/2-0.1$	$V_{CC}/2$	$V_{CC}/2+0.1$	V
$V_{REF(-)}$	Voltage, Bottom of Ladder	Measured at Ref(−)	−0.1	0		V
I_{IN}	Comparator Input Current	$f_c=640$ kHz, (Note 6)	−2	±0.5	2	μA

Electrical Characteristics

Digital Levels and DC Specifications: ADC0808CJ $4.5V\le V_{CC}\le 5.5V$, $-55°C\le T_A\le +125°C$ unless otherwise noted ADC0808CCJ, ADC0808CCN, ADC0808CCV, ADC0809CCN and ADC0809CCV, $4.75\le V_{CC}\le 5.25V$, $-40°C\le T_A\le +85°C$ unless otherwise noted

Symbol	Parameter	Conditions	Min	Typ	Max	Units
ANALOG MULTIPLEXER						
$I_{OFF(+)}$	OFF Channel Leakage Current	$V_{CC}=5V$, $V_{IN}=5V$,				
		$T_A=25°C$		10	200	nA
		T_{MIN} to T_{MAX}			1.0	μA
$I_{OFF(-)}$	OFF Channel Leakage Current	$V_{CC}=5V$, $V_{IN}=0$,				
		$T_A=25°C$	−200	−10		nA
		T_{MIN} to T_{MAX}	−1.0			μA

Electrical Characteristics (Continued)

Digital Levels and DC Specifications: ADC0808CJ $4.5V \le V_{CC} \le 5.5V$, $-55°C \le T_A \le +125°C$ unless otherwise noted ADC0808CCJ, ADC0808CCN, ADC0808CCV, ADC0809CCN and ADC0809CCV, $4.75 \le V_{CC} \le 5.25V$, $-40°C \le T_A \le +85°C$ unless otherwise noted

Symbol	Parameter	Conditions	Min	Typ	Max	Units
CONTROL INPUTS						
$V_{IN(1)}$	Logical "1" Input Voltage		$V_{CC}-1.5$			V
$V_{IN(0)}$	Logical "0" Input Voltage				1.5	V
$I_{IN(1)}$	Logical "1" Input Current (The Control Inputs)	$V_{IN}=15V$			1.0	μA
$I_{IN(0)}$	Logical "0" Input Current (The Control Inputs)	$V_{IN}=0$	−1.0			μA
I_{CC}	Supply Current	$f_{CLK}=640$ kHz		0.3	3.0	mA
DATA OUTPUTS AND EOC (INTERRUPT)						
$V_{OUT(1)}$	Logical "1" Output Voltage	$I_O=-360$ μA	$V_{CC}-0.4$			V
$V_{OUT(0)}$	Logical "0" Output Voltage	$I_O=1.6$ mA			0.45	V
$V_{OUT(0)}$	Logical "0" Output Voltage EOC	$I_O=1.2$ mA			0.45	V
I_{OUT}	TRI-STATE Output Current	$V_O=5V$ $V_O=0$	 −3		3	μA μA

Electrical Characteristics

Timing Specifications $V_{CC}=V_{REF(+)}=5V$, $V_{REF(-)}=GND$, $t_r=t_f=20$ ns and $T_A=25°C$ unless otherwise noted.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
t_{WS}	Minimum Start Pulse Width	*(Figure 5)*		100	200	ns
t_{WALE}	Minimum ALE Pulse Width	*(Figure 5)*		100	200	ns
t_s	Minimum Address Set-Up Time	*(Figure 5)*		25	50	ns
t_H	Minimum Address Hold Time	*(Figure 5)*		25	50	ns
t_D	Analog MUX Delay Time From ALE	$R_S=0\Omega$ *(Figure 5)*		1	2.5	μS
t_{H1}, t_{H0}	OE Control to Q Logic State	$C_L=50$ pF, $R_L=10$k *(Figure 8)*		125	250	ns
t_{1H}, t_{0H}	OE Control to Hi-Z	$C_L=10$ pF, $R_L=10$k *(Figure 8)*		125	250	ns
t_c	Conversion Time	$f_c=640$ kHz, *(Figure 5)* (Note 7)	90	100	116	μS
f_c	Clock Frequency		10	640	1280	kHz
t_{EOC}	EOC Delay Time	*(Figure 5)*	0		8 + 2 μS	Clock Periods
C_{IN}	Input Capacitance	At Control Inputs		10	15	pF
C_{OUT}	TRI-STATE Output Capacitance	At TRI-STATE Outputs, (Note 12)		10	15	pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to GND, unless othewise specified.

Note 3: A zener diode exists, internally, from V_{CC} to GND and has a typical breakdown voltage of 7 V_{DC}.

Note 4: Two on-chip diodes are tied to each analog input which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. The spec allows 100 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 100 mV, the output code will be correct. To achieve an absolute $0V_{DC}$ to $5V_{DC}$ input voltage range will therefore require a minimum supply voltage of 4.900 V_{DC} over temperature variations, initial tolerance and loading.

Note 5: Total unadjusted error includes offset, full-scale, linearity, and multiplexer errors. See *Figure 3*. None of these A/Ds requires a zero or full-scale adjust. However, if an all zero code is desired for an analog input other than 0.0V, or if a narrow full-scale span exists (for example: 0.5V to 4.5V full-scale) the reference voltages can be adjusted to achieve this. See *Figure 13*.

Note 6: Comparator input current is a bias current into or out of the chopper stabilized comparator. The bias current varies directly with clock frequency and has little temperature dependence *(Figure 8)*. See paragraph 4.0.

Note 7: The outputs of the data register are updated one clock cycle before the rising edge of EOC.

Note 8: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Functional Description

Multiplexer. The device contains an 8-channel single-ended analog signal multiplexer. A particular input channel is selected by using the address decoder. Table I shows the input states for the address lines to select any channel. The address is latched into the decoder on the low-to-high transition of the address latch enable signal.

TABLE I

SELECTED ANALOG CHANNEL	ADDRESS LINE C	B	A
IN0	L	L	L
IN1	L	L	H
IN2	L	H	L
IN3	L	H	H
IN4	H	L	L
IN5	H	L	H
IN6	H	H	L
IN7	H	H	H

CONVERTER CHARACTERISTICS

The Converter

The heart of this single chip data acquisition system is its 8-bit analog-to-digital converter. The converter is designed to give fast, accurate, and repeatable conversions over a wide range of temperatures. The converter is partitioned into 3 major sections: the 256R ladder network, the successive approximation register, and the comparator. The converter's digital outputs are positive true.

The 256R ladder network approach *(Figure 1)* was chosen over the conventional R/2R ladder because of its inherent monotonicity, which guarantees no missing digital codes. Monotonicity is particularly important in closed loop feedback control systems. A non-monotonic relationship can cause oscillations that will be catastrophic for the system. Additionally, the 256R network does not cause load variations on the reference voltage.

The bottom resistor and the top resistor of the ladder network in *Figure 1* are not the same value as the remainder of the network. The difference in these resistors causes the output characteristic to be symmetrical with the zero and full-scale points of the transfer curve. The first output transition occurs when the analog signal has reached $+\frac{1}{2}$ LSB and succeeding output transitions occur every 1 LSB later up to full-scale.

The successive approximation register (SAR) performs 8 iterations to approximate the input voltage. For any SAR type converter, n-iterations are required for an n-bit converter. *Figure 2* shows a typical example of a 3-bit converter. In the ADC0808, ADC0809, the approximation technique is extended to 8 bits using the 256R network.

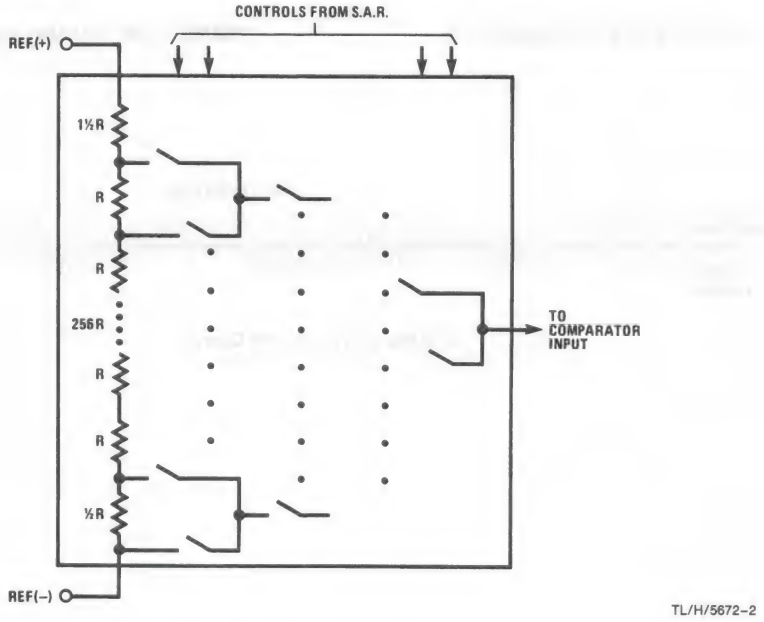

FIGURE 1. Resistor Ladder and Switch Tree

Functional Description (Continued)

The A/D converter's successive approximation register (SAR) is reset on the positive edge of the start conversion (SC) pulse. The conversion is begun on the falling edge of the start conversion pulse. A conversion in process will be interrupted by receipt of a new start conversion pulse. Continuous conversion may be accomplished by tying the end-of-conversion (EOC) output to the SC input. If used in this mode, an external start conversion pulse should be applied after power up. End-of-conversion will go low between 0 and 8 clock pulses after the rising edge of start conversion.

The most important section of the A/D converter is the comparator. It is this section which is responsible for the ultimate accuracy of the entire converter. It is also the comparator drift which has the greatest influence on the repeatability of the device. A chopper-stabilized comparator provides the most effective method of satisfying all the converter requirements.

The chopper-stabilized comparator converts the DC input signal into an AC signal. This signal is then fed throught a high gain AC amplifier and has the DC level restored. This technique limits the drift component of the amplifier since the drift is a DC component which is not passed by the AC amplifier. This makes the entire A/D converter extremely insensitive to temperature, long term drift and input offset errors.

Figure 4 shows a typical error curve for the ADC0808 as measured using the procedures outlined in AN-179.

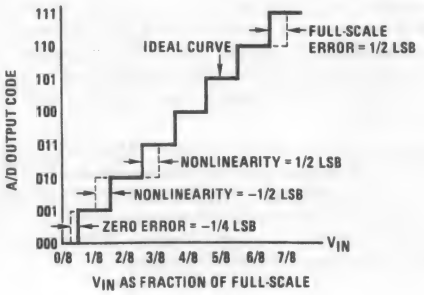

FIGURE 2. 3-Bit A/D Transfer Curve

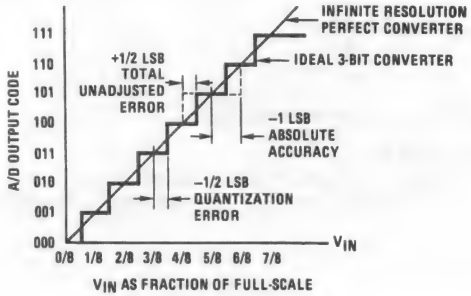

FIGURE 3. 3-Bit A/D Absolute Accuracy Curve

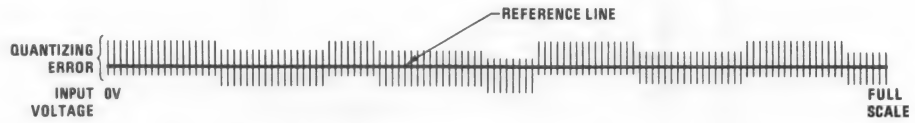

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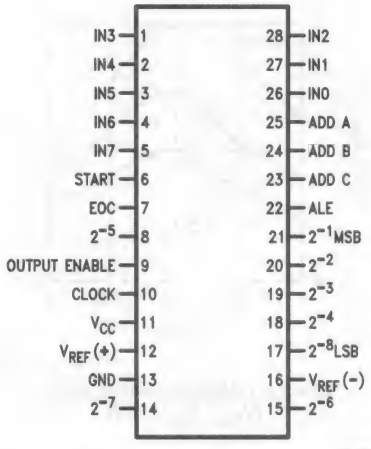
FIGURE 4. Typical Error Curve

Connection Diagrams

Dual-In-Line Package



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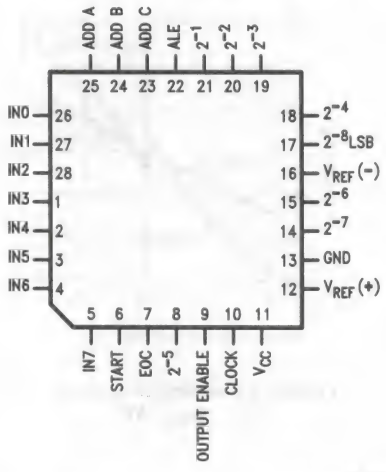
Order Number ADC0808CCN, ADC0809CCN, ADC0808CCJ or ADC0808CJ
See NS Package J28A or N28A

Molded Chip Carrier Package



TL/H/5672–12

Order Number ADC0808CCV or ADC0809CCV
See NS Package V28A

Timing Diagram

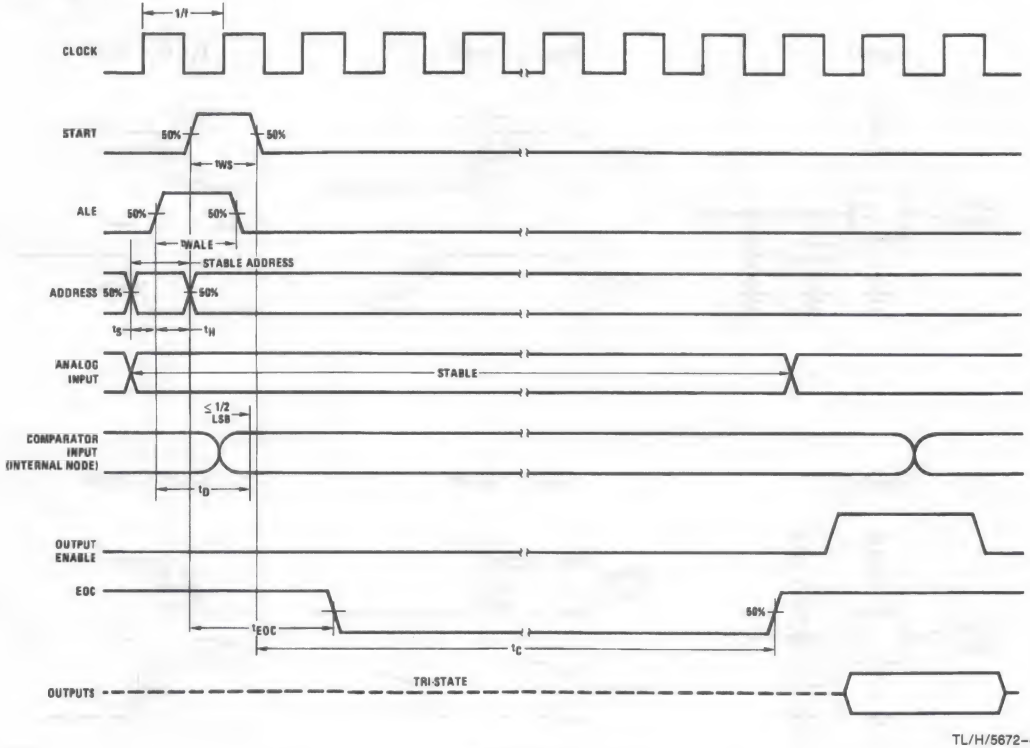


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FIGURE 5

Typical Performance Characteristics

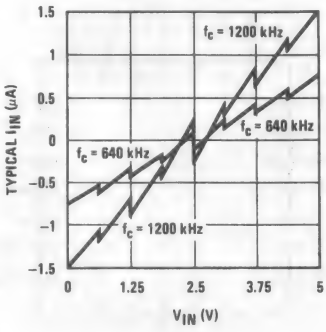
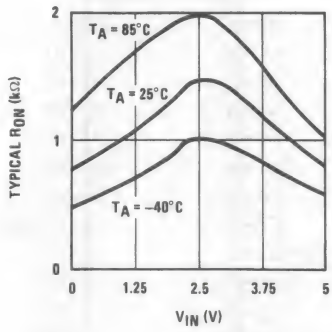


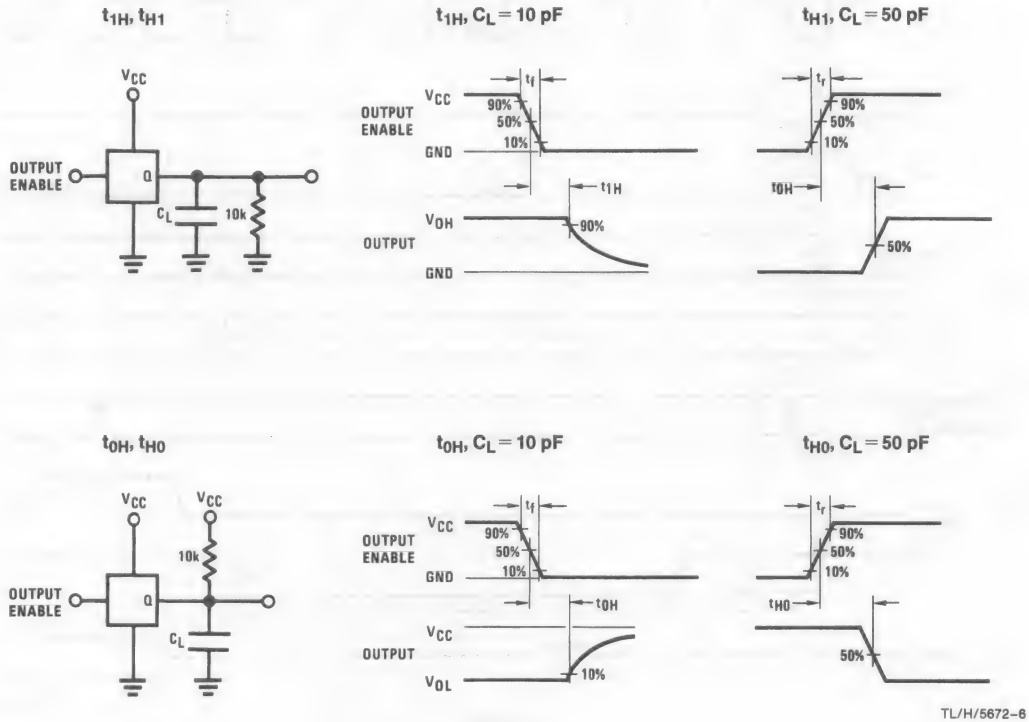
FIGURE 6. Comparator I_{IN} vs V_{IN}
($V_{CC} = V_{REF} = 5V$)



TL/H/5672–5

FIGURE 7. Multiplexer R_{ON} vs V_{IN}
($V_{CC} = V_{REF} = 5V$)

TRI-STATE Test Circuits and Timing Diagrams



TL/H/5672–6

FIGURE 8

Applications Information

OPERATION

1.0 RATIOMETRIC CONVERSION

The ADC0808, ADC0809 is designed as a complete Data Acquisition System (DAS) for ratiometric conversion systems. In ratiometric systems, the physical variable being measured is expressed as a percentage of full-scale which is not necessarily related to an absolute standard. The voltage input to the ADC0808 is expressed by the equation

$$\frac{V_{IN}}{V_{fs}-V_Z}=\frac{D_X}{D_{MAX}-D_{MIN}} \quad (1)$$

V_{IN} = Input voltage into the ADC0808

V_{fs} = Full-scale voltage

V_Z = Zero voltage

D_X = Data point being measured

D_{MAX} = Maximum data limit

D_{MIN} = Minimum data limit

A good example of a ratiometric transducer is a potentiometer used as a position sensor. The position of the wiper is directly proportional to the output voltage which is a ratio of the full-scale voltage across it. Since the data is represented as a proportion of full-scale, reference requirements are greatly reduced, eliminating a large source of error and cost for many applications. A major advantage of the ADC0808, ADC0809 is that the input voltage range is equal to the supply range so the transducers can be connected directly across the supply and their outputs connected directly into the multiplexer inputs, *(Figure 9)*.

Ratiometric transducers such as potentiometers, strain gauges, thermistor bridges, pressure transducers, etc., are suitable for measuring proportional relationships; however, many types of measurements must be referred to an absolute standard such as voltage or current. This means a system reference must be used which relates the full-scale voltage to the standard volt. For example, if $V_{CC}=V_{REF}=5.12V$, then the full-scale range is divided into 256 standard steps. The smallest standard step is 1 LSB which is then 20 mV.

2.0 RESISTOR LADDER LIMITATIONS

The voltages from the resistor ladder are compared to the selected into 8 times in a conversion. These voltages are coupled to the comparator via an analog switch tree which is referenced to the supply. The voltages at the top, center and bottom of the ladder must be controlled to maintain proper operation.

The top of the ladder, Ref(+), should not be more positive than the supply, and the bottom of the ladder, Ref(−), should not be more negative than ground. The center of the ladder voltage must also be near the center of the supply because the analog switch tree changes from N-channel switches to P-channel switches. These limitations are automatically satisfied in ratiometric systems and can be easily met in ground referenced systems.

Figure 10 shows a ground referenced system with a separate supply and reference. In this system, the supply must be trimmed to match the reference voltage. For instance, if a 5.12V is used, the supply should be adjusted to the same voltage within 0.1V.

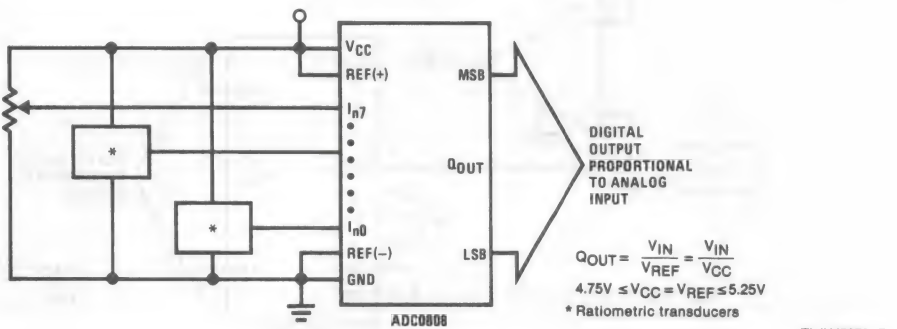


FIGURE 9. Ratiometric Conversion System

Applications Information (Continued)

The ADC0808 needs less than a milliamp of supply current so developing the supply from the reference is readily accomplished. In *Figure 11* a ground referenced system is shown which generates the supply from the reference. The buffer shown can be an op amp of sufficient drive to supply the milliamp of supply current and the desired bus drive, or if a capacitive bus is driven by the outputs a large capacitor will supply the transient supply current as seen in *Figure 12*. The LM301 is overcompensated to insure stability when loaded by the 10 μF output capacitor.

The top and bottom ladder voltages cannot exceed V_{CC} and ground, respectively, but they can be symmetrically less than V_{CC} and greater than ground. The center of the ladder voltage should always be near the center of the supply. The sensitivity of the converter can be increased, (i.e., size of the LSB steps decreased) by using a symmetrical reference system. In *Figure 13*, a 2.5V reference is symmetrically centered about $V_{CC}/2$ since the same current flows in identical resistors. This system with a 2.5V reference allows the LSB bit to be half the size of a 5V reference system.

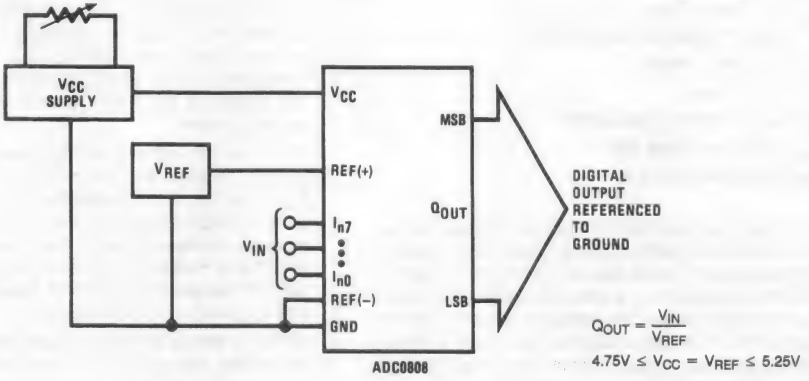


FIGURE 10. Ground Referenced Conversion System Using Trimmed Supply

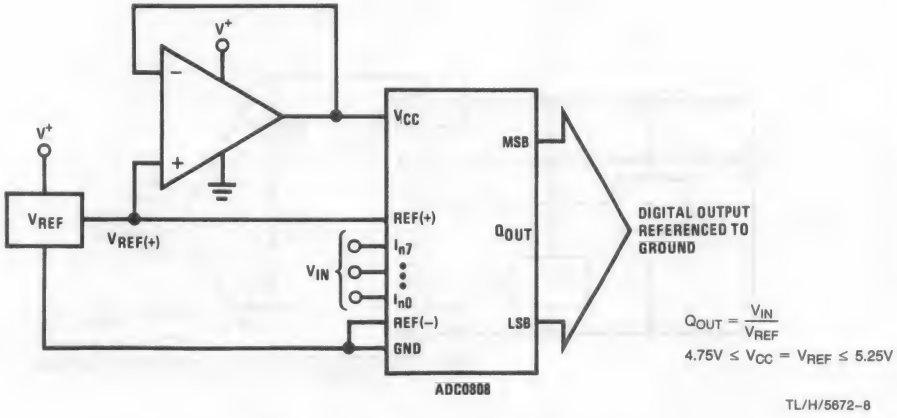


FIGURE 11: Ground Referenced Conversion System with Reference Generating V_{CC} Supply

Applications Information (Continued)

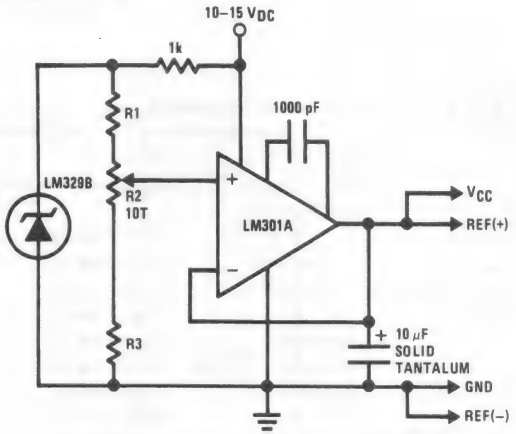


FIGURE 12. Typical Reference and Supply Circuit

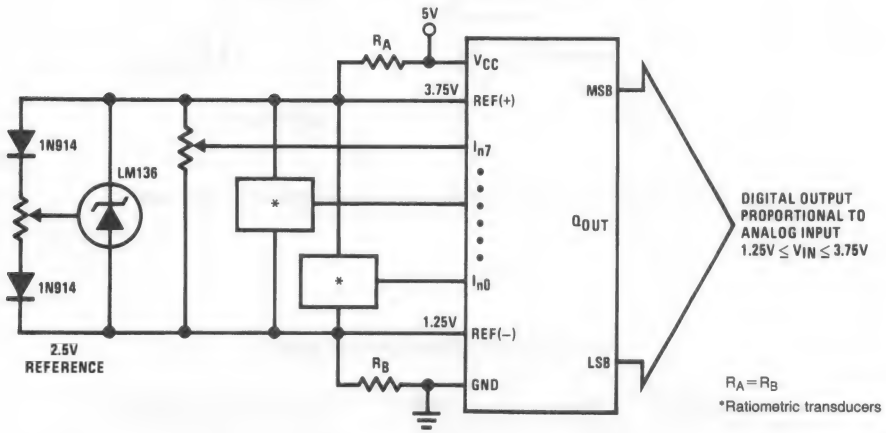


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FIGURE 13. Symmetrically Centered Reference

3.0 CONVERTER EQUATIONS

The transition between adjacent codes N and N+1 is given by:

$$V_{IN}=\left\{(V_{REF(+)}-V_{REF(-)})\left[\frac{N}{256}+\frac{1}{512}\right]\pm V_{TUE}\right\}+V_{REF(-)} \quad (2)$$

The center of an output code N is given by:

$$V_{IN}\left\{(V_{REF(+)}-V_{REF(-)})\left[\frac{N}{256}\right]\pm V_{TUE}\right\}+V_{REF(-)} \quad (3)$$

The output code N for an arbitrary input are the integers within the range:

$$N=\frac{V_{IN}-V_{REF(-)}}{V_{REF(+)}-V_{REF(-)}}\times 256\pm \text{Absolute Accuracy} \quad (4)$$

where: V_{IN} = Voltage at comparator input

$V_{REF(+)}$ = Voltage at Ref(+)

$V_{REF(-)}$ = Voltage at Ref(−)

V_{TUE} = Total unadjusted error voltage (typically $V_{REF(+)} \div 512$)

4.0 ANALOG COMPARATOR INPUTS

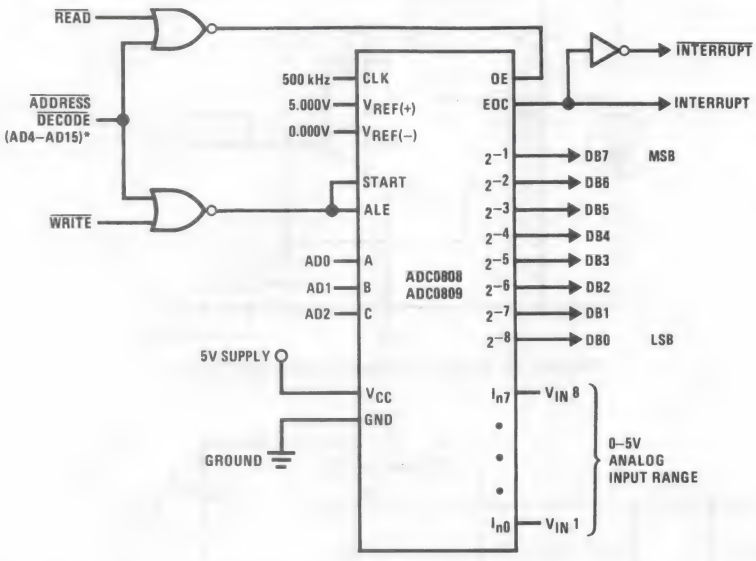
The dynamic comparator input current is caused by the periodic switching of on-chip stray capacitances. These are connected alternately to the output of the resistor ladder/switch tree network and to the comparator input as part of the operation of the chopper stabilized comparator.

The average value of the comparator input current varies directly with clock frequency and with V_{IN} as shown in *Figure 6.*

If no filter capacitors are used at the analog inputs and the signal source impedances are low, the comparator input current should not introduce converter errors, as the transient created by the capacitance discharge will die out before the comparator output is strobed.

If input filter capacitors are desired for noise reduction and signal conditioning they will tend to average out the dynamic comparator input current. It will then take on the characteristics of a DC bias current whose effect can be predicted conventionally.

Typical Application



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*Address latches needed for 8085 and SC/MP interfacing the ADC0808 to a microprocessor

MICROPROCESSOR INTERFACE TABLE

PROCESSOR	READ	WRITE	INTERRUPT (COMMENT)
8080	MEMR	MEMW	INTR (Thru RST Circuit)
8085	RD	WR	INTR (Thru RST Circuit)
Z-80	RD	WR	INT (Thru RST Circuit, Mode 0)
SC/MP	NRDS	NWDS	SA (Thru Sense A)
6800	VMA•ϕ2•R/W	VMA•ϕ•R/W	IRQA or IRQB (Thru PIA)

Ordering Information

TEMPERATURE RANGE		−40°C to +85°C			−55°C to +125°C
Error	±½ LSB Unadjusted	ADC0808CCN	ADC0808CCV	ADC0808CCJ	ADC0808CJ
	±1 LSB Unadjusted	ADC0809CCN	ADC0809CCV		
Package Outline		N28A Molded DIP	V28A Molded Chip Carrier	J28A Ceramic DIP	J28A Ceramic DIP

National Semiconductor

ADC0811 8-Bit Serial I/O A/D Converter With 11-Channel Multiplexer

General Description

The ADC0811 is an 8-Bit successive approximation A/D converter with simultaneous serial I/O. The serial input controls an analog multiplexer which selects from 11 input channels or an internal half scale test voltage.

An input sample-and-hold is implemented by a capacitive reference ladder and sampled data comparator. This allows the input signal to vary during the conversion cycle.

Separate serial I/O and conversion clock inputs are provided to facilitate the interface to various microprocessors.

Features

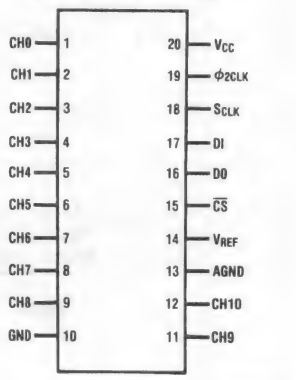
- Separate asynchronous converter clock and serial data I/O clock.
- 11-Channel multiplexer with 4-Bit serial address logic.
- Built-in sample and hold function.
- Ratiometric or absolute voltage referencing.
- No zero or full-scale adjust required.
- Internally addressable test voltage.
- 0V to 5V input range with single 5V power supply.
- TTL/MOS input/output compatible.
- 0.3″ standard width 20-pin dip or 20-pin molded chip carrier

Key Specifications

■ Resolution	8-Bits
■ Total unadjusted error	± ½LSB and ± 1LSB
■ Single supply	$5V_{DC}$
■ Low Power	15 mW
■ Conversion Time	32 μS

Connection Diagrams

Dual-In-Line Package



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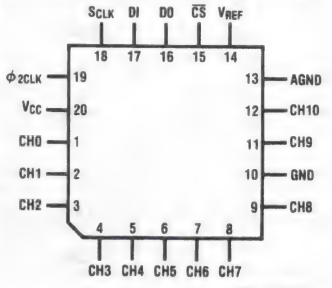
Top View

Molded Chip Carrier (PCC) Package

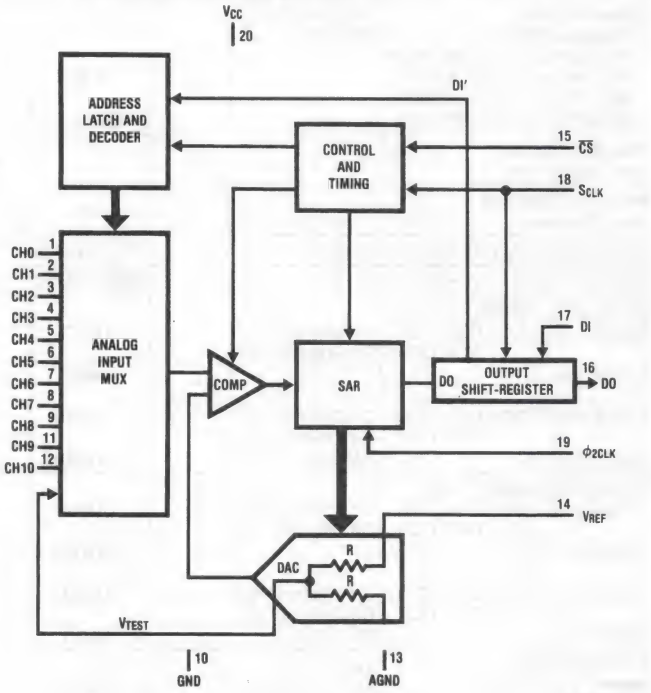


TL/H/5587–2

Top View

Order Number ADC0811J,N,V
See NS Packages J20A, N20A, V20A
Use Ordering Information

Functional Diagram



TL/H/5587–3

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	6.5V
Voltage	
Inputs and Outputs	−0.3V to V_{CC} +0.3V
Input Current Per Pin (Note 3)	±5mA
Total Package Input Current (Note 3)	±20mA
Storage Temperature	−65°C to +150°C
Package Dissipation at T_A = 25°C	875 mW
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C
ESD Susceptibility (Note 11)	2000V

Operating Ratings (Notes 1 & 2)

Supply Voltage (V_{CC})	4.5 V_{DC} to 6.0 V_{DC}
Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0811BCN, ADC0811CCN	$0°C \le T_A \le 70°C$
ADC0811BCJ, ADC0811BCV	$-40°C \le T_A \le 85°C$
ADC0811CCJ, ADC0811CCV	$-40°C \le T_A \le 85°C$
ADC0811BJ, ADC0811CJ	$-55°C \le T_A \le 125°C$

Electrical Characteristics

The following specifications apply for V_{CC} = 4.75V to 5.25V, V_{REF} = +4.6V to (V_{CC} + 0.1V), $\phi_{2\ CLK}$ = 2.097 MHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J$ = 25°C.

Parameter	Conditions	ADC0811BCJ, ADC0811BJ ADC0811CCJ, ADC0811CJ			ADC0811BCN, ADC0811BCV ADC0811CCN, ADC0811CCV			Units
		Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	
CONVERTER AND MULTIPLEXER CHARACTERISTICS								
Maximum Total Unadjusted Error	V_{REF} = 5.00 V_{DC} (Note 4)							
ADC0811BCN, ADC0811BCV						±½	±½	LSB
ADC0811BCJ, ADC0811BJ			±½					LSB
ADC0811CCN, ADC0811CCV						±1	±**1**	LSB
ADC0811CCJ, ADC0811CJ			±**1**					LSB
Minimum Reference Input Resistance		8		**5**	8		**5**	kΩ
Maximum Reference Input Resistance		8	**11**		8	11	**11**	kΩ
Maximum Analog Input Range	(Note 5)		**V_{CC} + 0.05**			V_{CC} + 0.05	**V_{CC} + 0.05**	V
Minimum Analog Input Range			**GND − 0.05**			GND − 0.05	**GND − 0.05**	V
On Channel Leakage Current								
ADC0811BCJ, CCJ, BCN, CCN, BCV, CCV	On Channel = 5V Off Channel = 0V		**1000**			400	**1000**	nA
ADC0811CJ, BJ			**1000**					nA
ADC0811BCJ, CCJ, BCN, CCN, BCV, CCV	On Channel = 0V Off Channel = 5V		**−1000**			−400	**−1000**	nA
ADC0811BJ, CJ	(Note 9)		**−1000**					nA
Off Channel Leakage Current								
ADC0811BCJ, CCJ, BCN, CCN, BCV, CCV	On Channel = 5V Off Channel = 0V		**−1000**			−400	**1000**	nA
ADC0811CJ, BJ			**−1000**					nA
ADC0811BCJ, CCJ, BCN, CCN, BCV, CCV	On Channel = 0V Off Channel = 5V		**1000**			400	**1000**	nA
ADC0811BJ, CJ	(Note 9)		**1000**					nA
Minimum V_{TEST} Internal Test Voltage	$V_{REF} = V_{CC}$, CH 11 Selected		**125**			125	**125**	(Note 10) Counts
Maximum V_{TEST} Internal Test Voltage	$V_{REF} = V_{CC}$, CH 11 Selected		**130**			130	**130**	(Note 10) Counts

Electrical Characteristics

The following specifications apply for V_{CC} = 4.75V to 5.25V, V_{REF} = +4.6V to (V_{CC} + 0.1V), $\phi_{2\,CLK}$ = 2.097 MHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J$ = 25°C. (Continued)

Parameter	Conditions	ADC0811BCJ, ADC0811BJ ADC0811CCJ, ADC0811CJ			ADC0811BCN, ADC0811BCV ADC0811CCN, ADC0811CCV			Units
		Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	
DIGITAL AND DC CHARACTERISTICS								
$V_{IN(1)}$, Logical "1" Input Voltage (Min)	V_{CC} = 5.25V		**2.0**			2.0	**2.0**	V
$V_{IN(0)}$, Logical "0" Input Voltage (Max)	V_{CC} = 4.75V		**0.8**			0.8	**0.8**	V
$I_{IN(1)}$, Logical "1" Input Current (Max)	V_{IN} = 5.0V	0.005	**2.5**		0.005	2.5	**2.5**	μA
$I_{IN(0)}$, Logical "0" Input Current (Max)	V_{IN} = 0V	−0.005	**−2.5**		−0.005	2.5	**−2.5**	μA
$V_{OUT(1)}$, Logical "1" Output Voltage (Min)	V_{CC} = 4.75V I_{OUT} = −360 μA I_{OUT} = −10 μA		**2.4** **4.5**			2.4 4.5	**2.4** **4.5**	V V
$V_{OUT(0)}$, Logical "0" Output Voltage (Max)	V_{CC} = 5.25V I_{OUT} = 1.6 mA		**0.4**			0.4	**0.4**	V
I_{OUT}, TRI-STATE Output Current (Max)	V_{OUT} = 0V V_{OUT} = 5V	−0.01 0.01	**−3** **3**		−0.01 0.01	−3 3	**−3** **3**	μA μA
I_{SOURCE}, Output Source Current (Min)	V_{OUT} = 0V	−12	**−6.5**		−14	−6.5	**−6.5**	mA
I_{SINK}, Output Sink Current (Min)	V_{OUT} = V_{CC}	18	**8.0**		16	8.0	**8.0**	mA
I_{CC}, Supply Current (Max)	$\overline{CS}$ = 1, V_{REF} Open	1	**2.5**		1	2.5	**2.5**	mA
I_{REF} (Max)	V_{REF} = 5V	**0.7**	**1**		**0.7**	1	**1**	mA

AC CHARACTERISTICS

Parameter		Conditions	Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
$\phi_{2\,CLK}$, ϕ_2 Clock Frequency	MIN		0.70		**1.0**	MHz
	MAX		3.0	2.0	**2.1**	
S_{CLK}, Serial Data Clock Frequency	MIN				**5.0**	KHz
	MAX		700	525	**525**	
T_C, Conversion Process Time	MIN	Not Including MUX Addressing and Analog Input Sampling Times	48		**48**	ϕ_2 cycles
	MAX		64		**64**	
t_{ACC}, Access Time Delay From $\overline{CS}$ Falling Edge to DO Data Valid	MIN				**1**	ϕ_2 cycles
	MAX				**3**	
t_{SET-UP}, Minimum Set-up Time of $\overline{CS}$ Falling Edge to S_{CLK} Rising Edge					$\mathbf{4/\phi_{2CLK} + \frac{1}{2\,S_{CLK}}}$	sec
$t_{H\overline{CS}}$, $\overline{CS}$ Hold Time After the Falling Edge of S_{CLK}					**0**	ns
$t_{\overline{CS}}$, Total $\overline{CS}$ Low Time	MIN				$\mathbf{t_{set\text{-}up} + 8/S_{CLK}}$	sec
	MAX				$\mathbf{t_{\overline{CS}}(min) + 48/\phi_{2CLK}}$	sec
t_{HDI}, Minimum DI Hold Time from S_{CLK} Rising Edge			0		**0**	ns
t_{HDO}, Minimum DO Hold Time from S_{CLK} Falling Edge		R_L = 30k, C_L = 100 pF			**10**	ns

Electrical Characteristics

The following specifications apply for V_{CC} = 4.75V to 5.25V, V_{REF} = +4.6V to (V_{CC} + 0.1V), $\phi_{2\ CLK}$ = 2.097 MHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits T_A = T_J = 25°C. (Continued)

Parameter	Conditions		Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
AC CHARACTERISTICS (Continued)						
t_{SDI}, Minimum DI Set-up Time to S_{CLK} Rising Edge			200		**400**	ns
t_{DDO}, Maximum Delay From S_{CLK} Falling Edge to DO Data Valid	R_L = 30k, C_L = 100 pF		180	400	**400**	ns
t_{TRI}, Maximum DO Hold Time, ($\overline{CS}$ Rising edge to DO TRI-STATE)	R_L = 3k, C_L = 100 pF		90	150	**150**	ns
t_{CA}, Analog Sampling Time	After Address Is Latched $\overline{CS}$ = Low				**4/S_{CLK} + 1 μs**	sec
t_{RDO}, Maximum DO	R_L = 30 kΩ,	"TRI-STATE" to "HIGH" State	75	150	**150**	ns
Rise Time	C_L = 100 pf	"LOW" to "HIGH" State	150	300	**300**	
t_{FDO}, Maximum DO	R_L = 30 kΩ,	"TRI-STATE" to "LOW" State	75	150	**150**	ns
Fall Time	C_L = 100 pf	"HIGH" to "LOW" State	150	300	**300**	
C_{IN}, Maximum Input	Analog Inputs, ANO–AN10 and V_{REF}		11		**55**	pF
Capacitance	All Others		5		**15**	

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to ground.

Note 3: Under over voltage conditions (V_{IN} < 0V and V_{IN} > V_{CC}) the maximum input current at any one pin is ±5 mA. If the voltage at more than one pin exceeds V_{CC} + .3V the total package current must be limited to 20 mA. For example the maximum number of pins that can be over driven at the maximum current level of ±5 mA is four.

Note 4: Total unadjusted error includes offset, full-scale, linearity, multiplexer, and hold step errors.

Note 5: Two on-chip diodes are tied to each analog input, which will forward-conduct for analog input voltages one diode drop below ground or one diode drop greater than V_{CC} supply. Be careful during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct, especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

Note 6: Typicals are at 25°C and represent most likely parametric norm.

Note 7: Guaranteed and 100% production tested under worst case condition.

Note 8: Guaranteed, but not 100% production tested. These limits are not used to calculate outgoing quality levels.

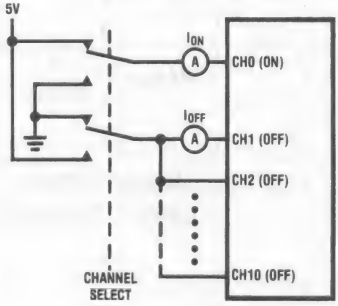
Note 9: Channel leakage current is measured after the channel selection.

Note 10: 1 count = V_{REF}/256.

Note 11: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Test Circuits

Leakage Current



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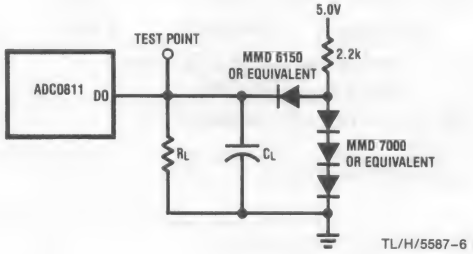
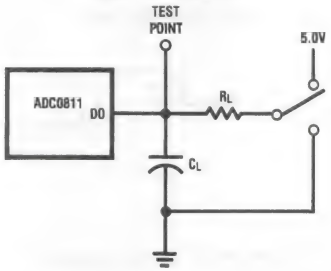
D0 Except "TRI-STATE"



TL/H/5587–6

Test Circuits (Continued)

t_{TRI} "TRI-STATE"



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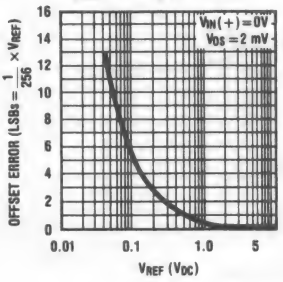
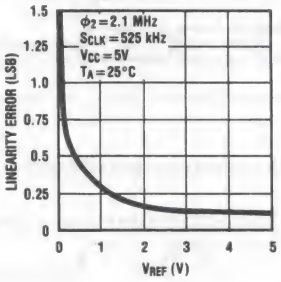
Typical Performance Characteristics

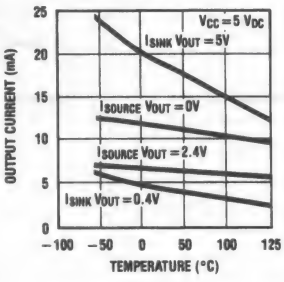
Unadjusted Offset Error vs V_{REF} Voltage



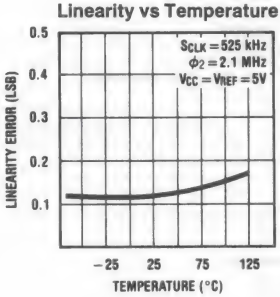
Linearity Error vs V_{REF} Voltage



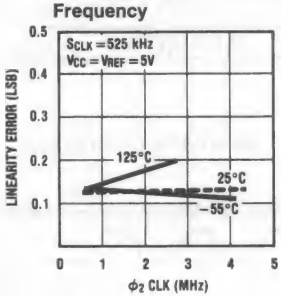
Output Current vs Temperature



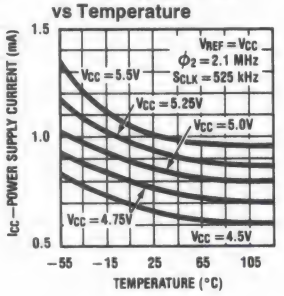
Linearity vs Temperature



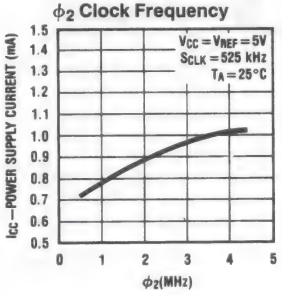
Linearity vs ϕ_2 Clock Frequency



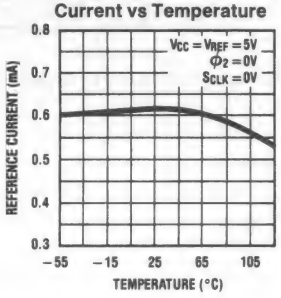
Power Supply Current vs Temperature



Power Supply Current vs ϕ_2 Clock Frequency



Resistive Ladder Reference Current vs Temperature



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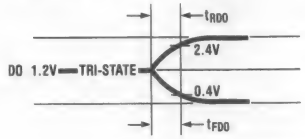
Timing Diagrams

D0 "TRI-STATE" Rise & Fall Times



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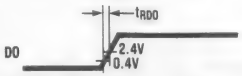
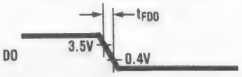
D0 Low to High State



TL/H/5587–14

D0 High to Low State



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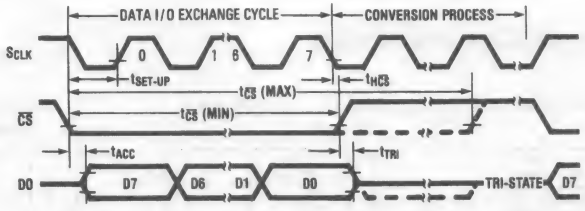
Timing with a continuous S_{CLK}



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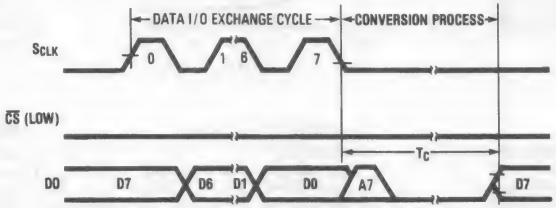
*Strobing $\overline{CS}$ High and Low will abort the present conversion and initiate a new serial I/O exchange.

Timing with a gated S_{CLK} and $\overline{CS}$ Continuously Low



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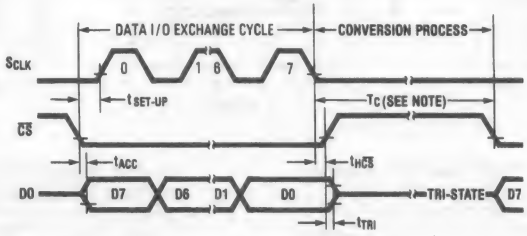
Using $\overline{CS}$ To TRI-STATE D0

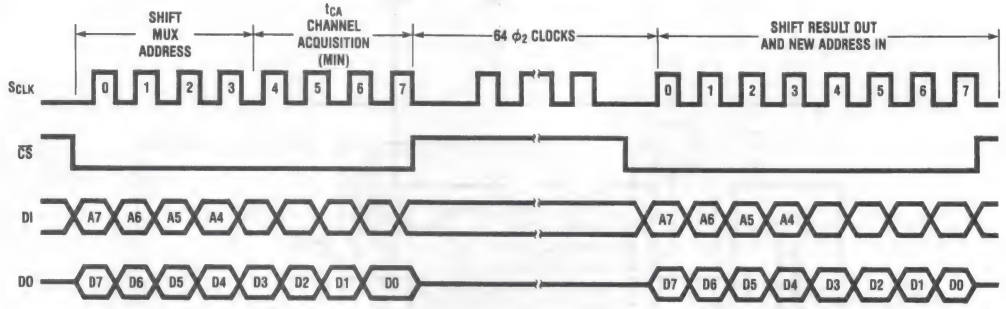


TL/H/5587–10

Note: Strobing $\overline{CS}$ Low during this time interval will abort the conversion in process.

Timing Diagrams (Continued)

$\overline{CS}$ High During Conversion



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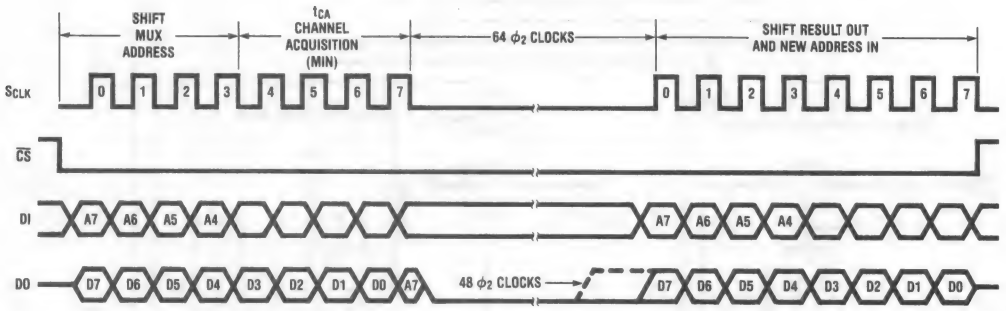
$\overline{CS}$ Low During Conversion



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Note: DO and DI lines share the 8-bit I/O shift register(see Functional Block Diagram). Since the MUX address bits are shifted in on S_{CLK} rising edges while S_{CLK} falling edges shift out conversion data on DO, the eighth falling edge of S_{CLK} will shift out the MSB MUX address bit (A7) on DO. Thus, if addressing channels CH8–CH10, a high DO will occur momentarily (one ϕ_2 clock period) until the 8-bit I/O shift register is cleared by the internal EOC signal.

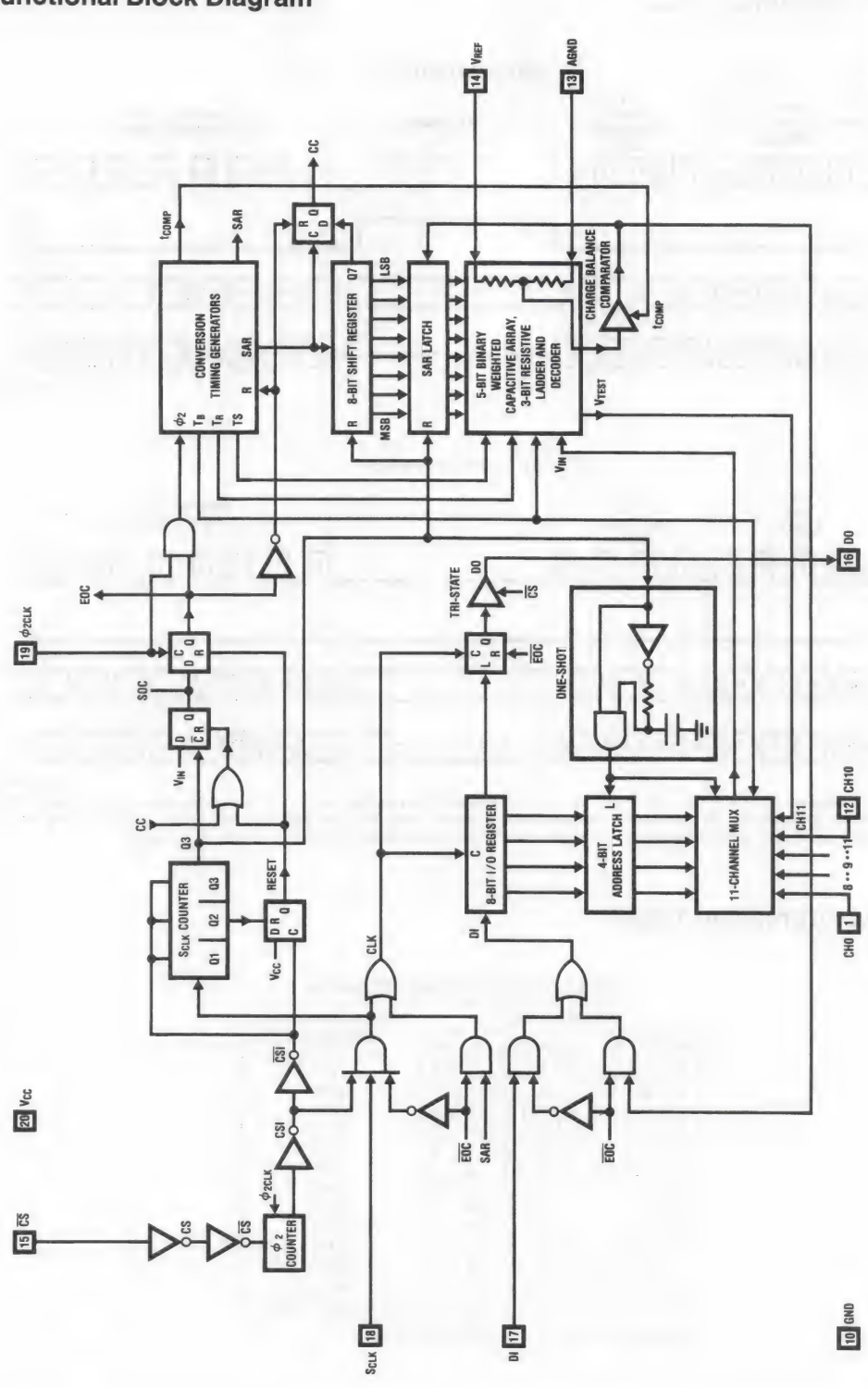
Channel Addressing Table

TABLE I. ADC 0811 Channel Addressing

MUX ADDRESS								ANALOG CHANNEL SELECTED
A_7	A_6	A_5	A_4	A_3	A_2	A_1	A_0	
0	0	0	0	X	X	X	X	CH0
0	0	0	1	X	X	X	X	CH1
0	0	1	0	X	X	X	X	CH2
0	0	1	1	X	X	X	X	CH3
0	1	0	0	X	X	X	X	CH4
0	1	0	1	X	X	X	X	CH5
0	1	1	0	X	X	X	X	CH6
0	1	1	1	X	X	X	X	CH7
1	0	0	0	X	X	X	X	CH8
1	0	0	1	X	X	X	X	CH9
1	0	1	0	X	X	X	X	CH10
1	0	1	1	X	X	X	X	V_{TEST}
1	1	X	X	X	X	X	X	LOGIC TEST MODE*

* Analog channel inputs CH0 thru CH3 are logic outputs

Functional Block Diagram



Functional Description

1.0 DIGITAL INTERFACE

The ADC0811 uses five input/output pins to implement the serial interface. Taking chip select ($\overline{CS}$) low enables the I/O data lines (DO and DI) and the serial clock input (S_{CLK}). The result of the last conversion is transmitted by the A/D on the DO line, while simultaneously the DI line receives the address data that selects the mux channel for the next conversion. The mux address is shifted in on the rising edge of S_{CLK} and the conversion data is shifted out on the falling edge. It takes eight S_{CLK} cycles to complete the serial I/O. A second clock (ϕ_2) controls the SAR during the conversion process and must be continuously enabled.

1.1 CONTINUOUS S_{CLK}

With a continuous S_{CLK} input $\overline{CS}$ must be used to synchronize the serial data exchange (see *Figure 1*). The ADC0811 recognizes a valid $\overline{CS}$ one to three ϕ_2 clock periods after the actual falling edge of $\overline{CS}$. This is implemented to ensure noise immunity of the $\overline{CS}$ signal. Any spikes on $\overline{CS}$ less than one ϕ_2 clock period will be ignored. $\overline{CS}$ must remain low during the complete I/O exchange which takes eight S_{CLK} cycles. Although $\overline{CS}$ is not immediately acknowledged for the purpose of starting a new conversion, the falling edge of $\overline{CS}$ immediately enables DO to output the MSB (D7) of the previous conversion.

The first S_{CLK} rising edge will be acknowledged after a set-up time ($t_{set\text{-}up}$) has elapsed from the falling edge of $\overline{CS}$. This and the following seven S_{CLK} rising edges will shift in the channel address for the analog multiplexer. Since there are 12 channels only four address bits are utilized. The first four S_{CLK} cycles clock in the mux address, during the next four S_{CLK} cycles the analog input is selected and sampled. During this mux address/sample cycle, data from the last conversion is also clocked out on DO. Since D7 was clocked out on the falling edge of $\overline{CS}$ only data bits D6–D0 remain to be received. The following seven falling edges of S_{CLK} shift out this data on DO.

The 8th S_{CLK} falling edge initiates the beginning of the A/D's actual conversion process which takes between 48 to 64 ϕ_2 cycles (T_C). During this time $\overline{CS}$ can go high to TRI-STATE DO and disable the S_{CLK} input or it can remain low. If $\overline{CS}$ is held low a new I/O exchange will not start until the conversion sequence has been completed, however once the conversion ends serial I/O will immediately begin. Since there is an ambiguity in the conversion time (T_C) synchronizing the data exchange is impossible. Therefore $\overline{CS}$ should go high before the 48th ϕ_2 clock has elasped and return low after the 64th ϕ_2 to synchronize serial communication.

A conversion or I/O operation can be aborted at any time by strobing $\overline{CS}$. If $\overline{CS}$ is high or low less than one ϕ_2 clock it will be ignored by the A/D. If the $\overline{CS}$ is strobed high or low between 1 to 3 ϕ_2 clocks the A/D may or may not respond. Therefore $\overline{CS}$ must be strobed high or low greater than 3 ϕ_2 clocks to ensure recognition. If a conversion or I/O exchange is aborted while in process the consequent data output will be erroneous until a complete conversion sequence has been implemented.

1.2 DISCONTINUOUS S_{CLK}

Another way to accomplish synchronous serial communication is to tie $\overline{CS}$ low continuously and disable S_{CLK} after its 8th falling edge (see *Figure 2*). S_{CLK} must remain low for

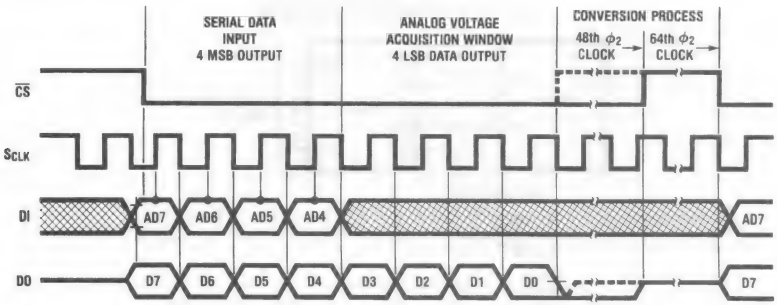


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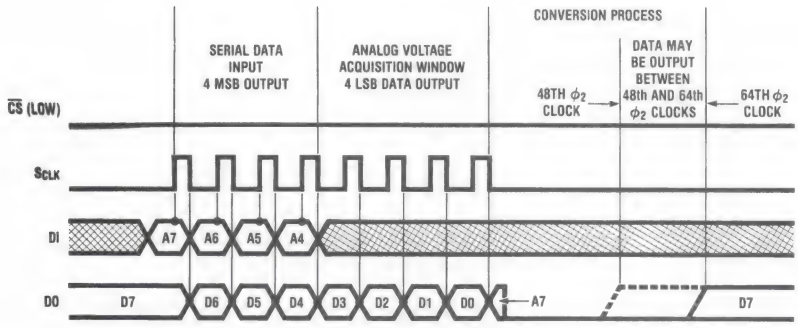
FIGURE 1



TL/H/5587–19

FIGURE 2

Functional Description (Continued)

at least 64 ϕ_2 clocks to insure that the A/D has completed its conversion. If S_{CLK} is enabled sooner, synchronizing to the data output on DO is not possible since an end of conversion signal from the A/D is not available and the actual conversion time is not known. With $\overline{CS}$ low during the conversion time (64 ϕ_2 max) DO will go low after the eighth falling edge of S_{CLK} and remain low until the conversion is completed. Once the conversion is through DO will transmit the MSB. The rest of the data will be shifted out once S_{CLK} is enabled as discussed previously.

If $\overline{CS}$ goes high during the conversion sequence DO is tri-stated, and the result is not affected so long as $\overline{CS}$ remains high until the end of the conversion.

1.2 MULTIPLEXER ADDRESSING

The four bit mux address is shifted, MSB first, into DI. Input data corresponds to the channel selected as shown in table 1. Care should be taken not to send an address greater than or equal to twelve (11XX) as this puts the A/D in a digital testing mode. In this mode the analog inputs CH0 thru CH3 become digital outputs, for our use in production testing.

2.0 ANALOG INPUT

2.1 THE INPUT SAMPLE AND HOLD

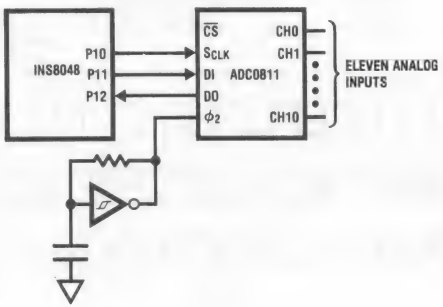
The ADC0811's sample/hold capacitor is implemented in its capacitive ladder structure. After the channel address is received, the ladder is switched to sample the proper analog input. This sampling mode is maintained for 1 μsec after the eighth S_{CLK} falling edge. The hold mode is initiated with the start of the conversion process. An acquisition window of $4t_{S_{CLK}}+1$ μsec is therefore available to allow the ladder capacitance to settle to the analog input voltage. Any change in the analog voltage before or after the acquisition window will not effect the A/D conversion result.

In the most simple case, the ladder's acquisition time is determined by the R_{on} (3K) of the multiplexer switches and the total ladder capacitance (90pf). These values yield an acquisition time of about 2 μsec for a full scale reading. Therefore the analog input must be stable for at least 2 μsec before and 1 μsec after the eighth S_{CLK} falling edge to ensure a proper conversion. External input source resistance and capacitance will lengthen the acquisition time and should be accounted for.

Other conventional sample and hold error specifications are included in the error and timing specs of the A/D. The hold step and gain error sample/hold specs are taken into account in the ADC0811's total unadjusted error, while the hold settling time is included in the A/D's max conversion time of 64 ϕ_2 clock periods. The hold droop rate can be thought of as being zero since an unlimited amount of time can pass between a conversion and the reading of data. However, once the data is read it is lost and another conversion is started.

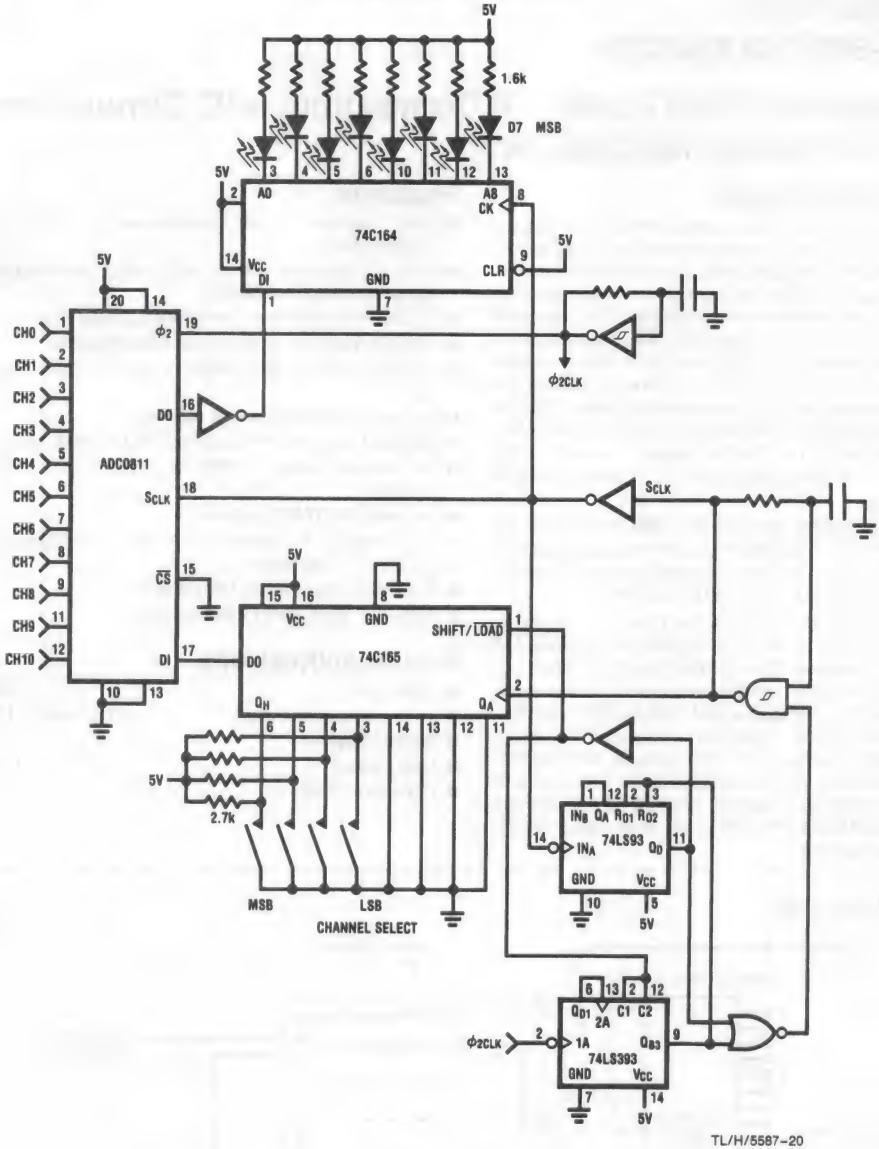
Typical Applications

ADC0811-INS8048 INTERFACE



TL/H/5587–21

ADC0811 FUNCTIONAL CIRCUIT



TL/H/5587–20

Ordering Information

Temperature Range		0°C to 70°C	−40°C to +85°C	−55°C to +125°C
Total Unadjusted Error	±½ LSB	ADC0811BCN	ADC0811BCJ ADC0811BCV	ADC0811BJ
	±1 LSB	ADC0811CCN	ADC0811CCJ ADC0811CCV	ADC0811CJ
Package Outline		N20A	J20A, V20A	J20A

National Semiconductor

ADC0816/ADC0817 8-Bit μP Compatible A/D Converters with 16-Channel Multiplexer

General Description

The ADC0816, ADC0817 data acquisition component is a monolithic CMOS device with an 8-bit analog-to-digital converter, 16-channel multiplexer and microprocessor compatible control logic. The 8-bit A/D converter uses successive approximation as the conversion technique. The converter features a high impedance chopper stabilized comparator, a 256R voltage divider with analog switch tree and a successive approximation register. The 16-channel multiplexer can directly access any one of 16-single-ended analog signals, and provides the logic for additional channel expansion. Signal conditioning of any analog input signal is eased by direct access to the multiplexer output, and to the input of the 8-bit A/D converter.

The device eliminates the need for external zero and full-scale adjustments. Easy interfacing to microprocessors is provided by the latched and decoded multiplexer address inputs and latched TTL TRI-STATE® outputs.

The design of the ADC0816, ADC0817 has been optimized by incorporating the most desirable aspects of several A/D conversion techniques. The ADC0816, ADC0817 offers high speed, high accuracy, minimal temperature dependence, excellent long-term accuracy and repeatability, and consumes minimal power. These features make this device ideally suited to applications from process and machine control to consumer and automotive applications. For similar performance in an 8-channel, 28-pin, 8-bit A/D converter, see the ADC0808, ADC0809 data sheet. (See AN-258 for more information.)

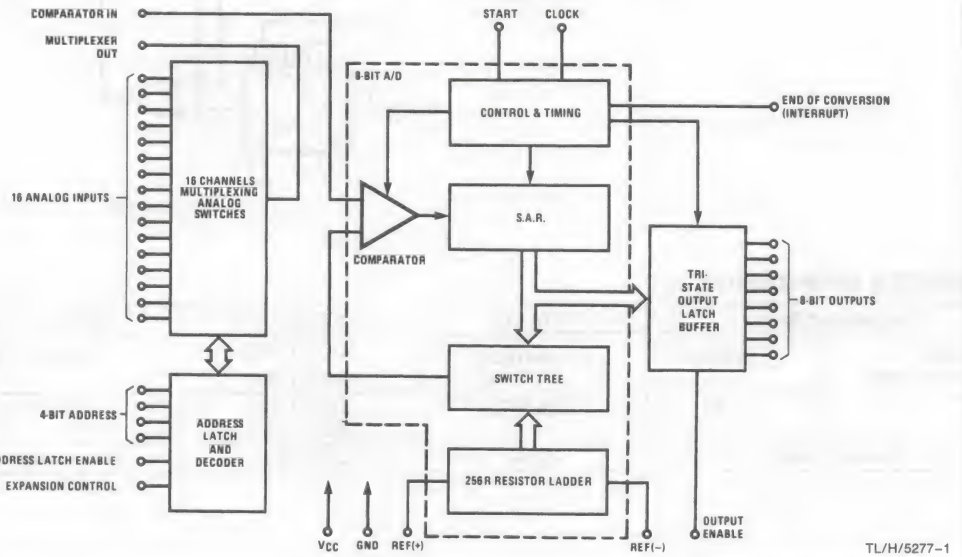
Features

- Easy interface to all microprocessors, or operates "stand alone"
- Operates ratiometrically or with 5 V_{DC} or analog span adjusted voltage reference
- 16-channel multiplexer with latched control logic
- Outputs meet TTL voltage level specifications
- 0V to 5V analog input voltage range with single 5V supply
- No zero or full-scale adjust required
- Standard hermetic or molded 40-pin DIP package
- Temperature range −40°C to +85°C or −55°C to +125°C
- Latched TRI-STATE output
- Direct access to "comparator in" and "multiplexer out" for signal conditioning
- ADC0816 equivalent to MM74C948
- ADC0817 equivalent to MM74C948-1

Key Specifications

■ Resolution	8 Bits
■ Total Unadjusted Error	±½ LSB and ±1 LSB
■ Single Supply	5 V_{DC}
■ Low Power	15 mW
■ Conversion Time	100 μs

Block Diagram



TL/H/5277–1

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC}) (Note 3)	6.5V
Voltage at Any Pin Except Control Inputs	−0.3V to (V_{CC}+0.3V)
Voltage at Control Inputs (START, OE, CLOCK, ALE, EXPANSION CONTROL, ADD A, ADD B, ADD C, ADD D)	−0.3V to 15V
Storage Temperature Range	−65°C to + 150°C
Package Dissipation at T_A = 25°C	875 mW
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
ESD Susceptibility (Note 9)	400V

Operating Conditions (Notes 1 & 2)

Temperature Range (Note 1)	$T_{MIN} \le T_A \le T_{MAX}$
ADC0816CJ	−55°C≤T_A+125°C
ADC0816CCJ, ADC0816CCN, ADC0817CCN	−40°C≤T_A≤+85°C
Range of V_{CC} (Note 1)	4.5 V_{DC} to 6.0 V_{DC}
Voltage at Any Pin Except Control Inputs	0V to V_{CC}
Voltage at Control Inputs (START, OE, CLOCK, ALE, EXPANSION CONTROL, ADD A, ADD B, ADD C, ADD D)	0V to 15V

Electrical Characteristics

Converter Specifications: V_{CC}=5 V_{DC}= $V_{REF(+)}$, $V_{REF(-)}$=GND, V_{IN}=$V_{COMPARATOR\ IN}$, $T_{MIN} \le T_{MAX}$ and f_{CLK} = 640 kHz unless otherwise stated.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
	ADC0816 Total Unadjusted Error (Note 5)	25°C T_{MIN} to T_{MAX}			±½ ±¾	LSB LSB
	ADC0817 Total Unadjusted Error (Note 5)	0°C to 70°C T_{MIN} to T_{MAX}			±1 ±1¼	LSB LSB
	Input Resistance	From Ref(+) to Ref(−)	1.0	4.5		kΩ
	Analog Input Voltage Range	(Note 4)V(+) or V(−)	GND−0.10		V_{CC}+0.10	V_{DC}
$V_{REF(+)}$	Voltage, Top of Ladder	Measured at Ref(+)		V_{CC}	V_{CC}+0.1	V
$\frac{V_{REF(+)}+V_{REF(-)}}{2}$	Voltage, Center of Ladder		V_{CC}/2−0.1	V_{CC}/2	V_{CC}/2+0.1	V
$V_{REF(-)}$	Voltage, Bottom of Ladder	Measured at Ref(−)	−0.1	0		V
	Comparator Input Current	f_c=640 kHz, (Note 6)	−2	±0.5	2	μA

Electrical Characteristics

Digital Levels and DC Specifications: ADC0816CJ 4.5V≤V_{CC}≤5.5V, −55°C≤T_A≤+ 125°C unless otherwise noted.
ADC0816CCJ, ADC0816CCN, ADC0817CCN 4.75V≤V_{CC}≤5.25V, −40°C≤T_A≤+85°C unless otherwise noted.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
ANALOG MULTIPLEXER						
R_{ON}	Analog Multiplexer ON Resistance	(Any Selected Channel) T_A=25°C, R_L=10k T_A=85°C T_A=125°C		1.5	3 6 9	kΩ kΩ kΩ
ΔR_{ON}	ΔON Resistance Between Any 2 Channels	(Any Selected Channel) R_L=10k		75		Ω
I_{OFF+}	OFF Channel Leakage Current	V_{CC}=5V, V_{IN}=5V, T_A=25°C T_{MIN} to T_{MAX}		10	200 1.0	nA μA
$I_{OFF(-)}$	OFF Channel Leakage Current	V_{CC}=5V, V_{IN}=0, T_A=25°C T_{MIN} to T_{Max}	−200 −1.0			nA μA
CONTROL INPUTS						
$V_{IN(1)}$	Logical "1" Input Voltage		V_{CC}−1.5			V
$V_{IN(0)}$	Logical "0" Input Voltage				1.5	V

Electrical Characteristics (Continued)

Digital Levels and DC Specifications: ADC0816CJ—$4.5V \le V_{CC} \le 5.5V$, $-55°C \le T_A \le +125°C$ unless otherwise noted.
ADC0816CCJ, ADC0816CCN, ADC0817CCN—$4.75V \le V_{CC} \le 5.25V$, $-40°C \le T_A \le +85°C$ unless otherwise noted.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
CONTROL INPUTS (Continued)						
$I_{IN(1)}$	Logical "1" Input Current (The Control Inputs)	$V_{IN} = 15V$			1.0	μA
$I_{IN(0)}$	Logical "0" Input Current (The Control Inputs)	$V_{IN} = 0$	−1.0			μA
I_{CC}	Supply Current	$f_{CLK} = 640$ kHz		0.3	3.0	mA
DATA OUTPUTS AND EOC (INTERRUPT)						
$V_{OUT(1)}$	Logical "1" Output Voltage	$I_O = -360$ μA, $T_A = 85°C$ $I_O = -300$ μA, $T_A = 125°C$	$V_{CC} - 0.4$			V
$V_{OUT(0)}$	Logical "0" Output Voltage	$I_O = 1.6$ mA			0.45	V
$V_{OUT(0)}$	Logical "0" Output Voltage EOC	$I_O = 1.2$ mA			0.45	V
I_{OUT}	TRI-STATE Output Current	$V_O = V_{CC}$ $V_O = 0$	 −3.0		3.0	μA μA

Electrical Characteristics

Timing Specifications: $V_{CC} = V_{REF(+)} = 5V$, $V_{REF(-)} = GND$, $t_r = t_f = 20$ ns and $T_A = 25°C$ unless otherwise noted.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
t_{WS}	Minimum Start Pulse Width	*(Figure 5)* (Note 7)		100	200	ns
t_{WALE}	Minimum ALE Pulse Width	*(Figure 5)*		100	200	ns
t_S	Minimum Address Set-Up Time	*(Figure 5)*		25	50	ns
T_H	Minimum Address Hold Time	*(Figure 5)*		25	50	ns
t_D	Analog MUX Delay Time from ALE	$R_S = 0\Omega$ *(Figure 5)*		1	2.5	μS
t_{H1}, t_{H0}	OE Control to Q Logic State	$C_L = 50$ pF, $R_L = 10$k *(Figure 8)*		125	250	ns
t_{1H}, t_{0H}	OE Control to Hi-Z	$C_L = 10$ pF, $R_L = 10$k *(Figure 8)*		125	250	ns
t_C	Conversion Time	$f_c = 640$ kHz, *(Figure 5)* (Note 8)	90	100	116	μs
f_c	Clock Frequency		10	640	1280	kHz
t_{EOC}	EOC Delay Time	*(Figure 5)*	0		8 + 2μs	Clock Periods
C_{IN}	Input Capacitance	At Control Inputs		10	15	pF
C_{OUT}	TRI-STATE Output Capacitance	At TRI-STATE Outputs (Note 8)		10	15	pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to GND, unless otherwise specified.

Note 3: A zener diode exists, internally, from V_{CC} to GND and has a typical breakdown voltage of 7 V_{DC}.

Note 4: Two on-chip diodes are tied to each analog input which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. The spec allows 100 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 100 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.900 V_{DC} over temperature variations, initial tolerance and loading.

Note 5: Total unadjusted error includes offset, full-scale, and linearity errors. See *Figure 3*. None of these A/Ds requires a zero or full-scale adjust. However, if an all zero code is desired for an analog input other than 0.0V, or if a narrow full-scale span exists (for example: 0.5V to 4.5V full-scale) the reference voltages can be adjusted to achieve this. See *Figure 13*.

Note 6: Comparator input current is a bias current into or out of the chopper stabilized comparator. The bias current varies directly with clock frequency and has little temperature dependence *(Figure 6)*. See paragraph 4.0.

Note 7: If start pulse is asynchronous with converter clock or if $f_c > 640$ kHz, the minimum start pulse width is 8 clock periods plus 2 μs. For synchronous operation at $f_c \le 640$ kHz take start high within 100 ns of clock going low.

Note 8: The outputs of the data register are updated one clock cycle before the rising edge of EOC.

Note 9: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Functional Description

Multiplexer: The device contains a 16-channel single-ended analog signal multiplexer. A particular input channel is selected by using the address decoder. Table 1 shows the input states for the address line and the expansion control line to select any channel. The address is latched into the decoder on the low-to-high transition of the address latch enable signal.

TABLE 1

Selected Analog Channel	Address Line				Expansion Control
	D	C	B	A	
IN0	L	L	L	L	H
IN1	L	L	L	H	H
IN2	L	L	H	L	H
IN3	L	L	H	H	H
IN4	L	H	L	L	H
IN5	L	H	L	H	H
IN6	L	H	H	L	H
IN7	L	H	H	H	H
IN8	H	L	L	L	H
IN9	H	L	L	H	H
IN10	H	L	H	L	H
IN11	H	L	H	H	H
IN12	H	H	L	L	H
IN13	H	H	L	H	H
IN14	H	H	H	L	H
IN15	H	H	H	H	H
All Channels OFF	X	X	X	X	L

X = don't care

Additional single-ended analog signals can be multiplexed to the A/D converter by disabling all the multiplexer inputs using the expansion control. The additional external signals are connected to the comparator input and the device ground. Additional signal conditioning (i.e., prescaling, sample and hold, instrumentation amplification, etc.) may also be added between the analog input signal and the comparator input.

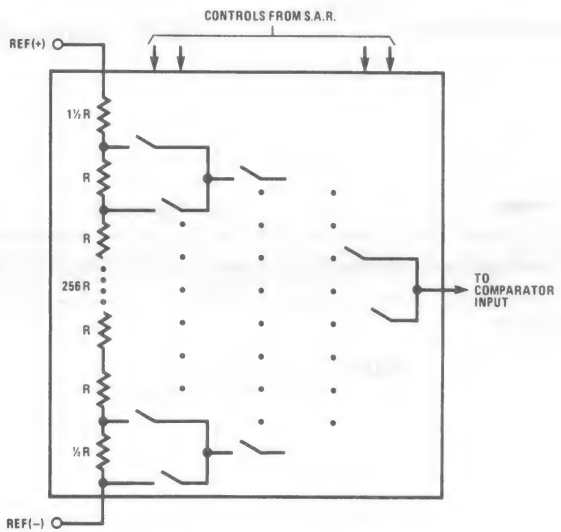
CONVERTER CHARACTERISTICS

The Converter

The heart of this single chip data acquisition system is its 8-bit analog-to-digital converter. The converter is designed to give fast, accurate, and repeatable conversions over a wide range of temperatures. The converter is partitioned into 3 major sections: the 256R ladder network, the successive approximation register, and the comparator. The converter's digital outputs are positive true.

The 256R ladder network approach *(Figure 1)* was chosen over the conventional R/2R ladder because of its inherent monotonicity, which guarantees no missing digital codes. Monotonicity is particularly important in closed loop feedback control systems. A non-monotonic relationship can cause oscillations that will be catastrophic for the system. Additionally, the 256R network does not cause load variations on the reference voltage.

The bottom resistor and the top resistor of the ladder network in *Figure 1* are not the same value as the remainder of the network. The difference in these resistors causes the output characteristic to be symmetrical with the zero and full-scale points of the transfer curve. The first output transition occurs when the analog signal has reached + ½ LSB and succeeding output transitions occur every 1 LSB later up to full-scale.



TL/H/5277–2

FIGURE 1. Resistor Ladder and Switch Tree

Functional Description (Continued)

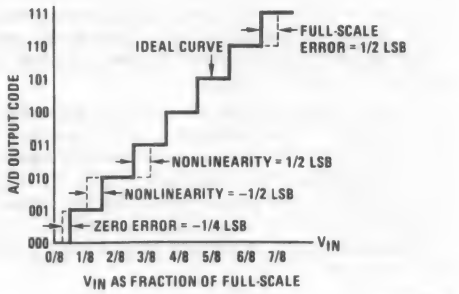
The successive approximation register (SAR) performs 8 iterations to approximate the input voltage. For any SAR type converter, n-iterations are required for an n-bit converter. *Figure 2* shows a typical example of a 3-bit converter. In the ADC0816, ADC0817, the approximation technique is extended to 8 bits using the 256R network.

The A/D converter's successive approximation register (SAR) is reset on the positive edge of the start conversion (SC) pulse. The conversion is begun on the falling edge of the start conversion pulse. A conversion in process will be interrupted by receipt of a new start conversion pulse. Continuous conversion may be accomplished by tying the end-of-conversion (EOC) output to the SC input. If used in this mode, an external start conversion pulse should be applied after power up. End-of-conversion will go low between 0 and 8 clock pulses after the rising edge of start conversion.

The most important section of the A/D converter is the comparator. It is this section which is responsible for the ulimate accuracy of the entire converter. It is also the comparator drift which has the greatest influence on the repeatability of the device. A chopper-stabilized comparator provides the most effective method of satisfying all the converter requirements.

The chopper-stabilized comparator converts the DC input signal into an AC signal. This signal is then fed through a high gain AC amplifier and has the DC level restored. This technique limits the drift component of the amplifier since the drift is a DC component which is not passed by the AC amplifier. This makes the entire A/D converter extremely insensitive to temperature, long term drift and input offset errors.

Figure 4 shows a typical error curve for the ADC0816 as measured using the procedures outlined in AN-179.



TL/H/5277–3

FIGURE 2. 3-Bit A/D Transfer Curve

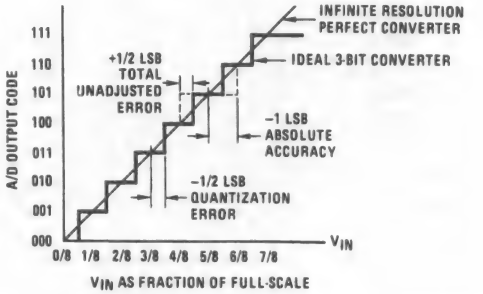


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FIGURE 3. 3-Bit A/D Absolute Accuracy Curve

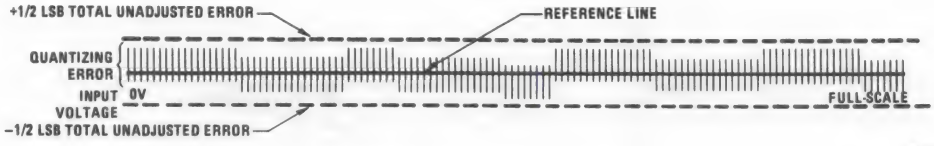

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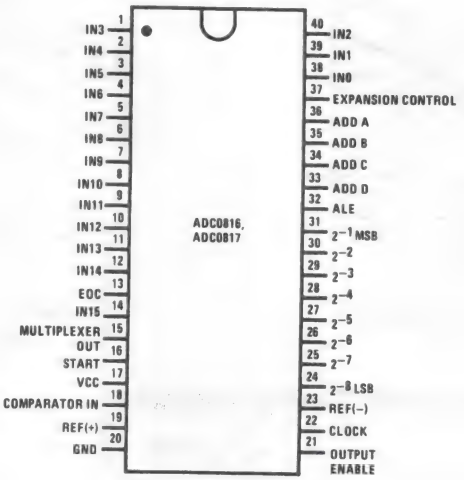
FIGURE 4. Typical Error Curve

Connection Diagram

Dual-In-Package



TOP VIEW

Order Number
ADC0816CCN,
ADC0817CCN,
ADC0816CCJ or
ADC0816CJ
See NS Package Number
J40A or N40A

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Timing Diagram

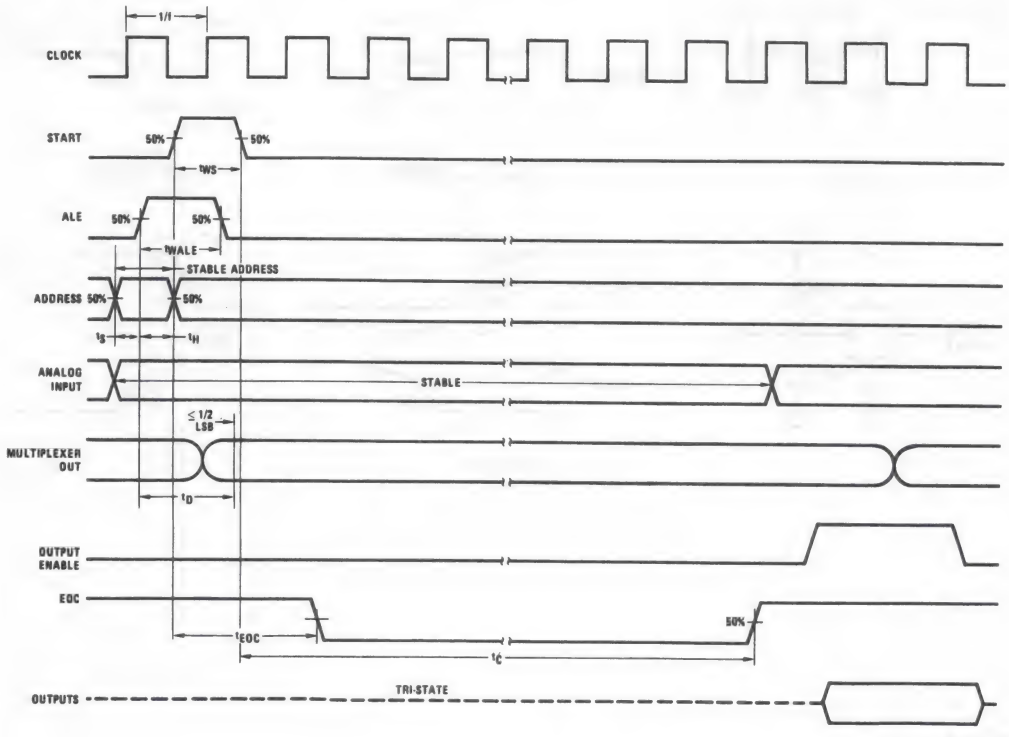

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FIGURE 5

Typical Performance Characteristics

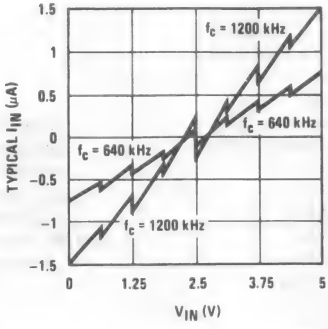

FIGURE 6. Comparator I_{IN} vs V_{IN} ($V_{CC} = V_{REF} = 5V$)

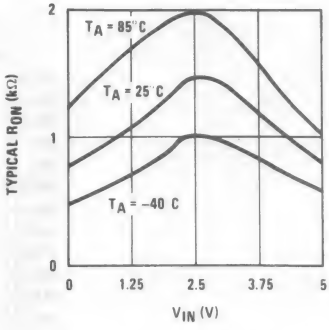

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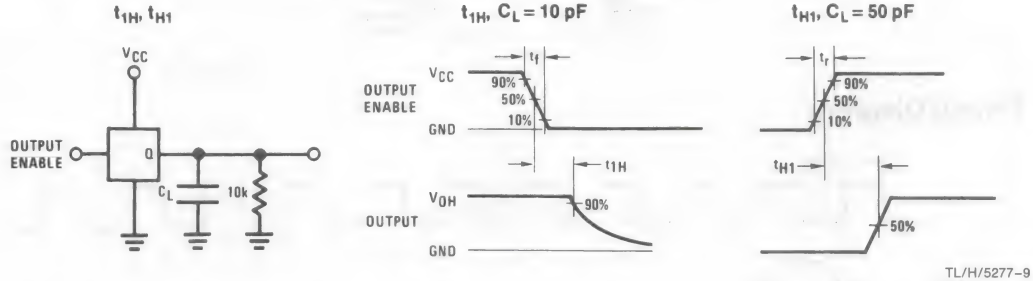
FIGURE 7. Multiplexer R_{ON} vs V_{IN} ($V_{CC} = V_{REF} = 5V$)

TRI-STATE Test Circuits and Timing Diagrams



TL/H/5277–9

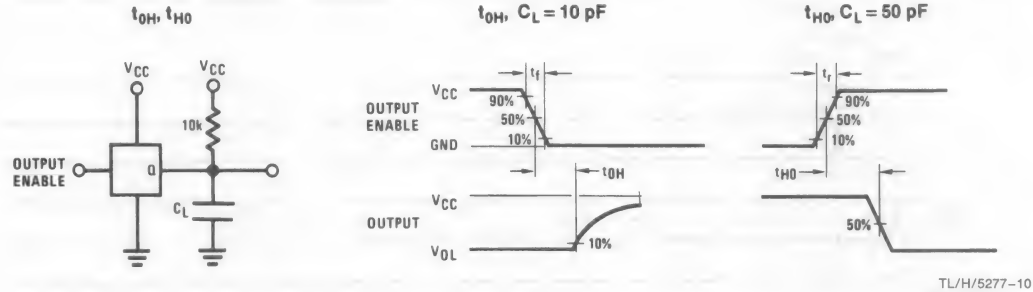


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FIGURE 8

Applications Information

OPERATION

1.0 RATIOMETRIC CONVERSION

The ADC0816, ADC0817 is designed as a complete Data Acquisition System (DAS) for ratiometric conversion systems. In ratiometric systems, the physical variable being measured is expressed as a percentage of full-scale which is not necessarily related to an absolute standard. The voltage input to the ADC0816 is expressed by the equation

$$\frac{V_{IN}}{V_{fs}-V_Z}=\frac{D_X}{D_{MAX}-D_{MIN}} \quad (1)$$

V_{IN} = Input voltage into the ADC0816
V_{fs} = Full-scale voltage
V_Z = Zero voltage
D_X = Data point being measured
D_{MAX} = Maximum data limit
D_{MIN} = Minimum data limit

A good example of a ratiometric transducer is a potentiometer used as a position sensor. The position of the wiper is directly proportional to the output voltage which is a ratio of the full-scale voltage across it. Since the data is represented as a proportion of full-scale, reference requirements are greatly reduced, eliminating a large source of error and cost for many applications. A major advantage of the ADC0816, ADC0817 is that the input voltage range is equal to the supply range so the transducers can be connected directly across the supply and their outputs connected directly into the multiplexer inputs, (*Figure 9*).

Ratiometric transducers such as potentiometers, strain gauges, thermistor bridges, pressure transducers, etc., are suitable for measuring proportional relationships; however, many types of measurements must be referred to an absolute standard such as voltage or current. This means a system reference must be used which relates the full-scale voltage to the standard volt. For example, if $V_{CC} = V_{REF} =$ 5.12V, then the full-scale range is divided into 256 standard steps. The smallest standard step is 1 LSB which is then 20 mV.

2.0 RESISTOR LADDER LIMITATIONS

The voltages from the resistor ladder are compared to the selected input 8 times in a conversion. These voltages are coupled to the comparator via an analog switch tree which is referenced to the supply. The voltages at the top, center and bottom of the ladder must be controlled to maintain proper operation.

The top of the ladder, Ref(+), should not be more positive than the supply, and the bottom of the ladder, Ref(−), should not be more negative than ground. The center of the ladder voltage must also be near the center of the supply because the analog switch tree changes from N-channel switches to P-channel switches These limitations are automaticaly satisfied in ratiometric systems and can be easily met in ground referenced systems.

Figure 10 shows a ground referenced system with a separate supply and reference. In this system, the supply must be trimmed to match the reference voltage. For instance, if a 5.12V reference is used, the supply should be adjusted to the same voltage within 0.1V.

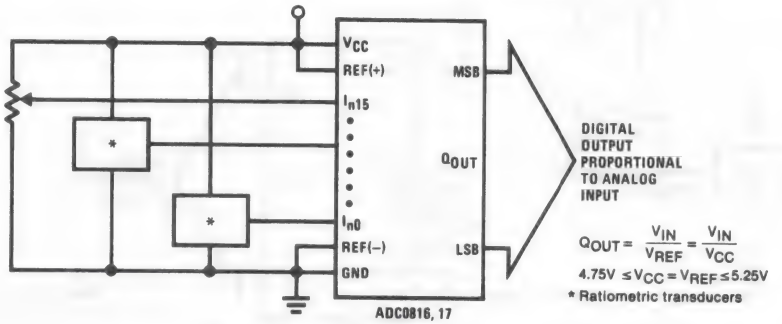

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FIGURE 9. Ratiometric Conversion System

Applications Information (Continued)

The ADC0816 needs less than a milliamp of supply current so developing the supply from the reference is readily accomplished. In *Figure 11* a ground references system is shown which generates the supply from the reference. The buffer shown can be an op amp of sufficient drive to supply the millliamp of supply current and the desired bus drive, or if a capacitive bus is driven by the outputs a large capacitor will supply the transient supply current as seen in *Figure 12*. The LM301 is overcompensated to insure stability when loaded by the 10 μF output capacitor.

The top and bottom ladder voltages cannot exceed V_{CC} and ground, respectively, but they can be symmetrically less than V_{CC} and greater than ground. The center of the ladder voltage should always be near the center of the supply. The sensitivity of the converter can be increased, (i.e., size of the LSB steps decreased) by using a symmetrical reference system. In *Figure 13*, a 2.5V reference is symmetrically centered about $V_{CC}/2$ since the same current flows in identical resistors. This system with a 2.5V reference allows the LSB to be half the size of the LSB in a 5V reference system.

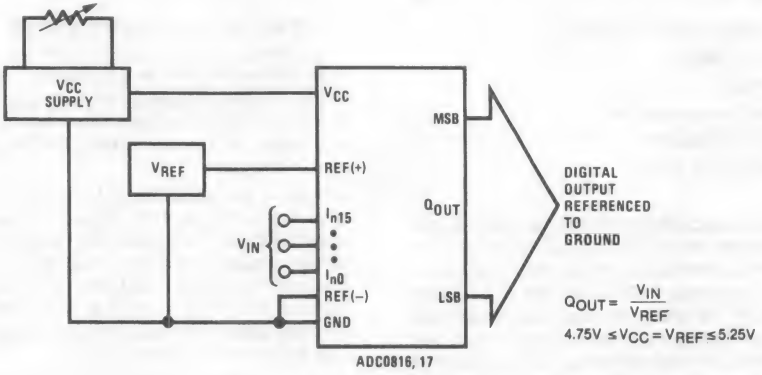


FIGURE 10. Ground Referenced Conversion System Using Trimmed Supply

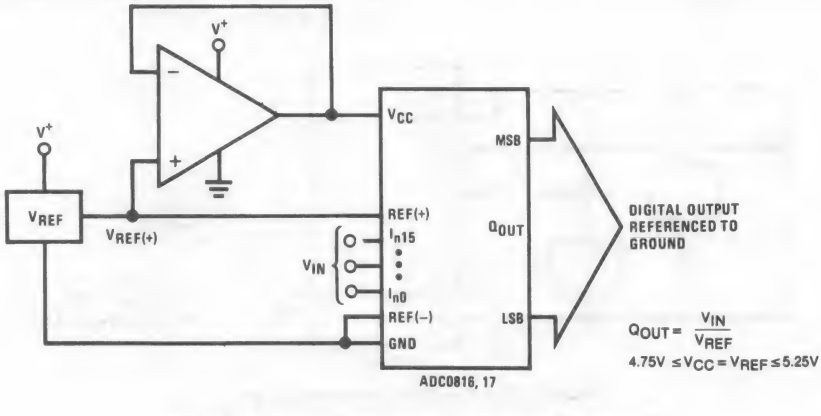
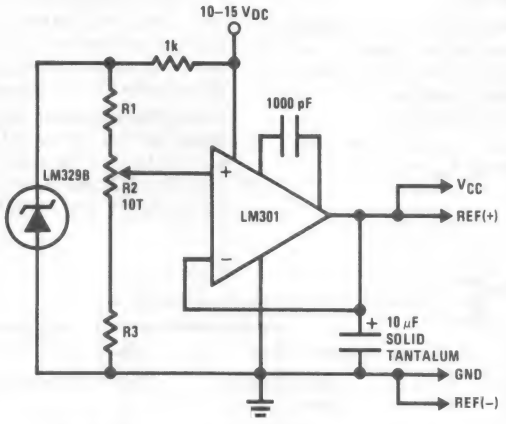


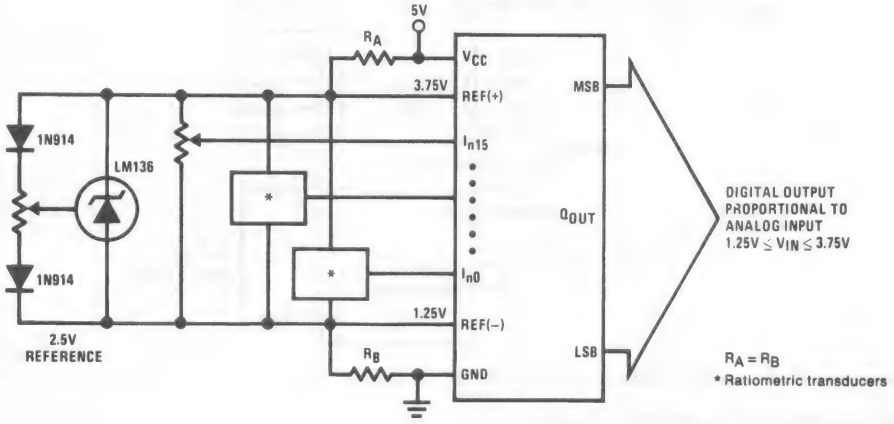
FIGURE 11. Ground Referenced Conversion System with Reference Generating V_{CC} Supply

Applications Information (Continued)



TL/H/5277–14

FIGURE 12. Typical Reference and Supply Circuit



TL/H/5277–15

FIGURE 13. Symmetrically Centered Reference

3.0 CONVERTER EQUATIONS

The transition between adjacent codes N and N + 1 is given by:

$$V_{IN} = \left\{ (V_{REF(+)} - V_{REF(-)}) \left[\frac{N}{256} + \frac{1}{512} \right] \pm V_{TUE} \right\} + V_{REF(-)} \quad (2)$$

The center of an output code N is given by:

$$V_{IN} = \left\{ (V_{REF(+)} - V_{REF(-)}) \left[\frac{N}{256} \right] \pm V_{TUE} \right\} + V_{REF(-)} \quad (3)$$

The output code N for an arbitrary input are the integers within the range:

$$N = \frac{V_{IN} - V_{REF(-)}}{V_{REF(+)} - V_{REF(-)}} \times 256 \pm \text{Absolute Accuracy} \quad (4)$$

where: V_{IN} = Voltage at comparator input

V_{REF} = Voltage at Ref(+)

V_{REF} = Voltage at Ref(−)

V_{TUE} = Total unadjusted error voltage (typically $V_{REF(+)} \div 512$)

Applications Information (Continued)

4.0 ANALOG COMPARATOR INPUTS

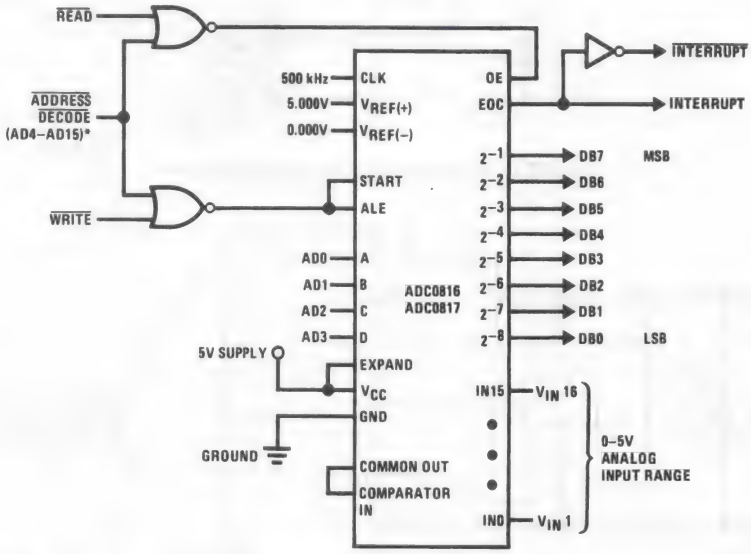
The dynamic comparator input current is caused by the periodic switching of on-chip stray capacitances These are connected alternately to the output of the resistor ladder/switch tree network and to the comparator input as part of the operation of the chopper stabilized comparator.

The average value of the comparator input current varies directly with clock frequency and with V_{IN} as shown in *Figure 6.*

If no filter capacitors are used at the analog or comparator inputs and the signal source impedances are low, the comparator input current should not introduce converter errors, as the transient created by the capacitance discharge will die out before the comparator output is strobed.

If input filter capacitors are desired for noise reduction and signal conditioning they will tend to average out the dynamic comparator input current. It will then take on the characteristics of a DC bias current whose effect can be predicted conventionally. See AN-258 for further discussion.

Typical Application



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*Address latches needed for 8085 and SC/MP interfacing the ADC0816, 17 to a microprocessor

Microprocessor Interface Table

PROCESSOR	$\overline{\text{READ}}$	$\overline{\text{WRITE}}$	INTERRUPT (COMMENT)
8080	$\overline{\text{MEMR}}$	$\overline{\text{MEMW}}$	INTR (Thru RST Circuit)
8085	$\overline{\text{RD}}$	$\overline{\text{WR}}$	INTR (Thru RST Circuit)
Z-80	$\overline{\text{RD}}$	$\overline{\text{WR}}$	$\overline{\text{INT}}$ (Thru RST Circuit, Mode 0)
SC/MP	NRDS	NWDS	SA (Thru Sense A)
6800	VMA•ϕ 2•R/W	VMA•Q_2•$\overline{\text{R/W}}$	$\overline{\text{IRQA}}$ or $\overline{\text{IRQB}}$ (Thru PIA)

Ordering Information

TEMPERATURE RANGE		−40°C to +85°C		−55°C to + 125°C
Error	±½ Bit Unadjusted	ADC0816CCN	ADC0816CCJ	ADC0816CJ
	±1 Bit Unadjusted	ADC0817CCN		
Package Outline		N40A Molded DIP	J40A Hermetic DIP	J40A Hermetic DIP

National Semiconductor

ADC0819 8-Bit Serial I/O A/D Converter with 19-Channel Multiplexer

General Description

The ADC0819 is an 8-Bit successive approximation A/D converter with simultaneous serial I/O. The serial input controls an analog multiplexer which selects from 19 input channels or an internal half scale test voltage.

An input sample-and-hold is implemented by a capacitive reference ladder and sampled data comparator. This allows the input signal to vary during the conversion cycle.

Separate serial I/O and conversion clock inputs are provided to facilitate the interface to various microprocessors.

Features

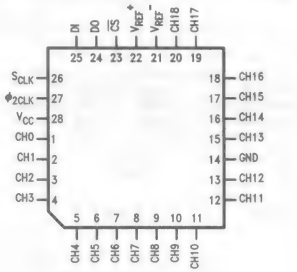
- Separate asynchronous converter clock and serial data I/O clock.
- 19-Channel multiplexer with 5-Bit serial address logic.
- Built-in sample and hold function.
- Ratiometric or absolute voltage referencing.
- No zero or full-scale adjust required.
- Internally addressable test voltage.
- 0V to 5V input range with single 5V power supply.
- TTL/MOS input/output compatible.
- 28-pin molded chip carrier or 28-pin molded DIP

Key Specifications

■ Resolution	8-Bits
■ Total unadjusted error	± ½LSB and ± 1LSB
■ Single supply	$5V_{DC}$
■ Low Power	15 mW
■ Conversion Time	16 μs

Connection Diagrams

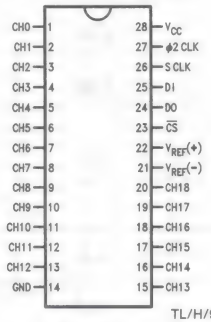
Molded Chip Carrier (PCC) Package



TL/H/9287–1

Top View
Order Number ADC0819BCV, CCV
See NS Package Number V28A

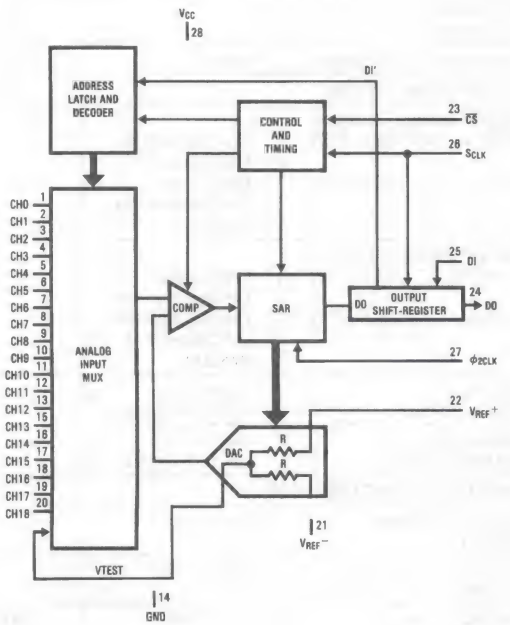
Dual-In-Line Package



TL/H/9287–20

Top View
Order Number ADC0819BCN, CCN
See NS Package Number N28B

Functional Diagram



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Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	6.5V
Voltage	
Inputs and Outputs	−0.3V to V_{CC} +0.3V
Input Current Per Pin (Note 3)	±5mA
Total Package Input Current (Note 3)	±20mA
Storage Temperature	−65°C to +150°C
Package Dissipation at T_A = 25°C	875 mW
Lead Temperature (Soldering, 10 sec.)	
Dual-In-Line Package (Plastic)	260°C
Surface Mount Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C
ESD Susceptibility (Note 11)	2000V

Operating Ratings (Notes 1 & 2)

Supply Voltage (V_{CC})	4.5 V_{DC} to 6.0 V_{DC}
Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0819BCV, ADC0819CCV	−40°C ≤ T_A ≤ +85°C
ADC0819BCN, ADC0819CCN	0°C ≤ T_A ≤ +70°C

Electrical Characteristics

The following specifications apply for V_{CC} = 5V, V_{REF} = 5V, $\phi_{2\,CLK}$ = 2.097 MHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J$ = 25°C.

Parameter	Conditions	ADC0819BCV, ADC0819BCN ADC0819CCV, ADC0819CCN			Units
		Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	
CONVERTER AND MULTIPLEXER CHARACTERISTICS					
Maximum Total Unadjusted Error ADC0819BCV, BCN ADC0819CCV, CCN	V_{REF} = 5.00 V_{DC} (Note 4)		 ±½ ±1	 **±½** **±1**	 LSB LSB
Minimum Reference Input Resistance		8		**5**	kΩ
Maximum Reference Input Resistance		8	11	**11**	kΩ
Maximum Analog Input Range	(Note 5)		V_{CC} + 0.05	**V_{CC} + 0.05**	V
Minimum Analog Input Range			GND − 0.05	**GND − 0.05**	V
On Channel Leakage Current ADC0819BCV, CCV, BCN, CCN ADC0819BCV, CCV, BCN, CCN	 On Channel = 5V Off Channel = 0V On Channel = 0V Off Channel = 5V (Note 9)		 400 −400	 **1000** **−1000**	 nA nA
Off Channel Leakage Current ADC0819BCV, CCV, BCN, CCN ADC0819BCV, CCV, BCN, CCN	 On Channel = 5V Off Channel = 0V On Channel = 0V Off Channel = 5V (Note 9)		 −400 400	 **−1000** **1000**	 nA nA
Minimum V_{TEST} Internal Test Voltage	V_{REF} = V_{CC}, CH 19 Selected		125	**125**	(Note 10) Counts
Maximum V_{TEST} Internal Test Voltage	V_{REF} = V_{CC}, CH 19 Selected		130	**130**	(Note 10) Counts
DIGITAL AND DC CHARACTERISTICS					
$V_{IN(1)}$, Logical "1" Input Voltage (Min)	V_{CC} = 5.25V		2.0	**2.0**	V
$V_{IN(0)}$, Logical "0" Input Voltage (Max)	V_{CC} = 4.75V		0.8	**0.8**	V
$I_{IN(1)}$, Logical "1" Input Current (Max)	V_{IN} = 5.0V	0.005	2.5	**2.5**	μA
$I_{IN(0)}$, Logical "0" Input Current (Max)	V_{IN} = 0V	−0.005	−2.5	**−2.5**	μA

Electrical Characteristics (Continued)

The following specifications apply for V_{CC} = 5V, V_{REF} = 5V, $\phi_{2\,CLK}$ = 2.097 MHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J$ = 25°C.

Parameter	Conditions	ADC0819BCV, ADC0819BCN ADC0819CCV, ADC0819CCN			Units
		Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	
DIGITAL AND DC CHARACTERISTICS (Continued)					
$V_{OUT(1)}$, Logical "1" Output Voltage (Min)	V_{CC}=4.75V I_{OUT}=−360 μA I_{OUT}=−10 μA		2.4 4.5	**2.4** **4.5**	V V
$V_{OUT(0)}$, Logical "0" Output Voltage (Max)	V_{CC}=5.25V I_{OUT}=1.6 mA		0.4	**0.4**	V
I_{OUT}, TRI-STATE Output Current (Max)	V_{OUT}=0V V_{OUT}=5V	−0.01 0.01	−3 3	**−3** **3**	μA μA
I_{SOURCE}, Output Source Current (Min)	V_{OUT}=0V	−14	−6.5	**−6.5**	mA
I_{SINK}, Output Sink Current (Min)	V_{OUT}=V_{CC}	16	8.0	**8.0**	mA
I_{CC}, Supply Current (Max)	$\overline{CS}$=1, V_{REF} Open	1	2.5	**2.5**	mA
I_{REF} (Max)	V_{REF}=5V	**0.7**	1	**1**	mA

AC CHARACTERISTICS

Parameter		Conditions	Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
$\phi_{2\,CLK}$, ϕ_2 Clock Frequency	MIN		0.70		**1.0**	MHz
	MAX		4.0	2.0	**2.1**	
S_{CLK}, Serial Data Clock Frequency	MIN				**5.0**	KHz
	MAX		1000	525	**525**	
T_C, Conversion Process Time	MIN	Not Including MUX Addressing and Analog Input Sampling Times	26		**26**	ϕ_2 cycles
	MAX		32		**32**	
t_{ACC}, Access Time Delay From $\overline{CS}$ Falling Edge to DO Data Valid	MIN				**1**	ϕ_2 cycles
	MAX				**3**	
t_{SET-UP}, Minimum Set-up Time of $\overline{CS}$ Falling Edge to S_{CLK} Rising Edge					$4/\phi_{2CLK} + \frac{1}{2\,S_{CLK}}$	sec
$t_{H\overline{CS}}$, $\overline{CS}$ Hold Time After the Falling Edge of S_{CLK}					**0**	ns
$t_{\overline{CS}}$, Total $\overline{CS}$ Low Time	MIN				$t_{set-up} + 8/S_{CLK}$	sec
	MAX				$t_{\overline{CS}}(min) + 26/\phi_{2CLK}$	sec
t_{HDI}, Minimum DI Hold Time from S_{CLK} Rising Edge			0		**0**	ns
t_{HDO}, Minimum DO Hold Time from S_{CLK} Falling Edge		R_L=30k, C_L=100 pF			**10**	ns
t_{SDI}, Minimum DI Set-up Time to S_{CLK} Rising Edge			200		**400**	ns
t_{DDO}, Maximum Delay From S_{CLK} Falling Edge to DO Data Valid		R_L=30k, C_L=100 pF	180	200	**250**	ns
t_{TRI}, Maximum DO Hold Time, ($\overline{CS}$ Rising edge to DO TRI-STATE)		R_L=3k, C_L=100 pF	90	150	**150**	ns

Electrical Characteristics

The following specifications apply for V_{CC} = 5V, $t_r = t_f = 20$ ns, V_{REF} = 5V, unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX}**; all other limits $T_A = T_J = 25°C$.

Parameter	Conditions		Typical (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
AC CHARACTERISTICS (Continued)						
t_{CA}, Analog Sampling Time	After Address Is Latched $\overline{CS}$ = Low				**$3/f_{CLK} + 1\ \mu s$**	sec
t_{RDO}, Maximum DO Rise Time	R_L = 30 kΩ, C_L = 100 pf	"TRI-STATE" to "HIGH" State	75	150	**150**	ns
		"LOW" to "HIGH" State	150	300	**300**	
t_{FDO}, Maximum DO Fall Time	R_L = 30 kΩ, C_L = 100 pf	"TRI-STATE" to "LOW" State	75	150	**150**	ns
		"HIGH" to "LOW" State	150	300	**300**	
C_{IN}, Maximum Input Capacitance	Analog Inputs, ANO–AN10 and V_{REF}		11		**55**	pF
	All Others		5		**15**	

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to ground.

Note 3: Under over voltage conditions ($V_{IN} < 0V$ and $V_{IN} > V_{CC}$) the maximum input current at any one pin is ±5 mA. If the voltage at more than one pin exceeds V_{CC} + .3V the total package current must be limited to 20 mA. For example the maximum number of pins that can be over driven at the maximum current level of ±5 mA is four.

Note 4: Total unadjusted error includes offset, full-scale, linearity, multiplexer, and hold step errors.

Note 5: Two on-chip diodes are tied to each analog input, which will forward-conduct for analog input voltages one diode drop below ground or one diode drop greater than V_{CC} supply. Be careful during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct, especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

Note 6: Typicals are at 25°C and represent most likely parametric norm.

Note 7: Tested Limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 8: Design Limits are guaranteed, but not 100% production tested. These limits are not used to calculate outgoing quality levels.

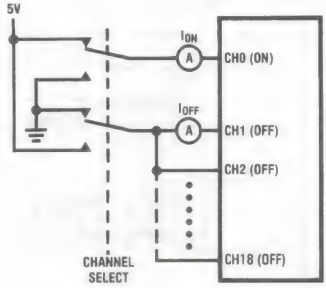
Note 9: Channel leakage current is measured after the channel selection.

Note 10: 1 count = $V_{REF}/256$.

Note 11: Human body model; 100 pF discharged through a 1.5 kΩ resistor.

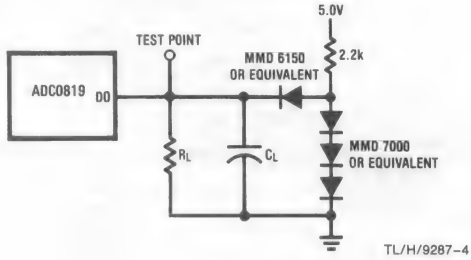
Test Circuits

Leakage Current



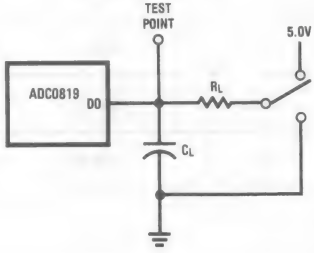
TL/H/9287–3

D0 Except "TRI-STATE"



TL/H/9287–4

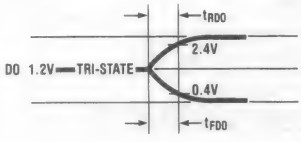
t_{TRI} "TRI-STATE"



TL/H/9287–5

Timing Diagrams

D0 "TRI-STATE" Rise & Fall Times



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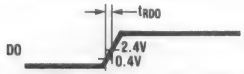
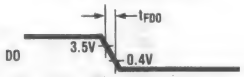
Timing Diagrams (Continued)

D0 Low to High State



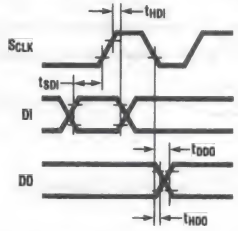
TL/H/9287–7

D0 High to Low State



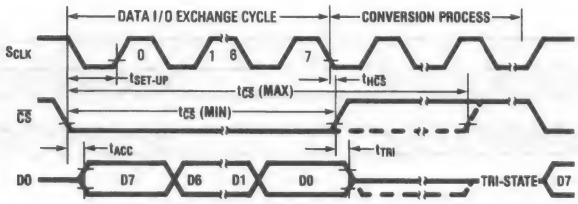
TL/H/9287–8

Data Input and Output Timing



TL/H/9287–9

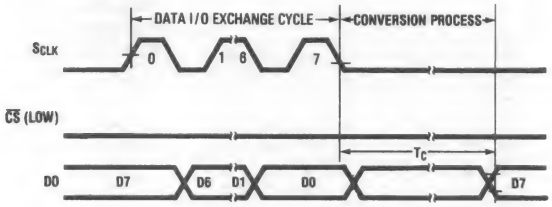
Timing with a continuous S_{CLK}



TL/H/9287–10

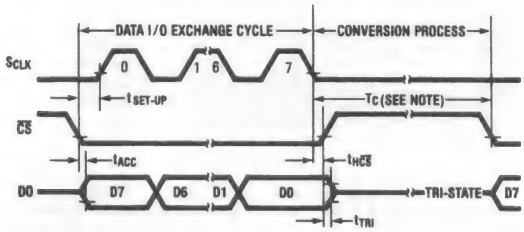
*Strobing $\overline{CS}$ High and Low will abort the present conversion and initiate a new serial I/O exchange.

Timing with a gated S_{CLK} and $\overline{CS}$ Continuously Low



TL/H/9287–11

Using $\overline{CS}$ To TRI-STATE D0

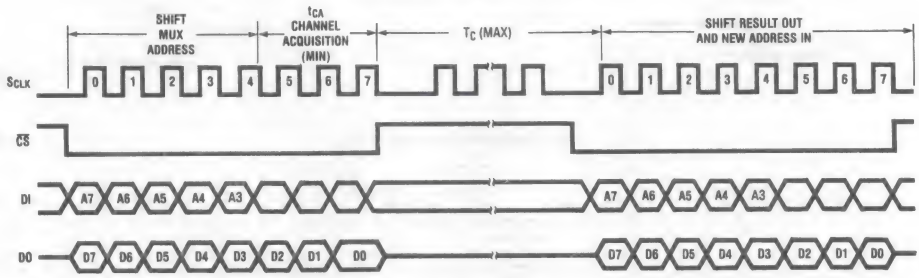


TL/H/9287–12

Note: Strobing $\overline{CS}$ Low during this time interval will abort the conversion in process.

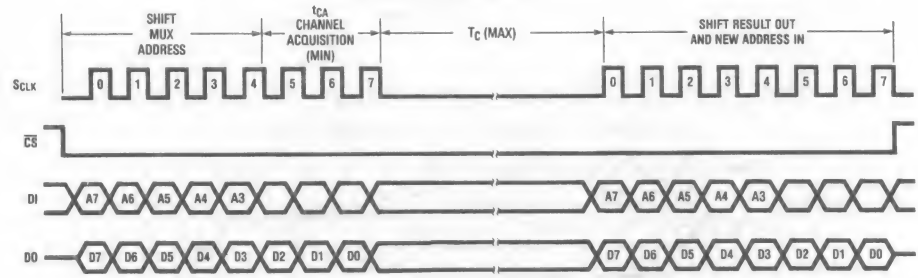
Timing Diagrams (Continued)

$\overline{CS}$ High During Conversion



TL/H/9287–13

$\overline{CS}$ Low During Conversion



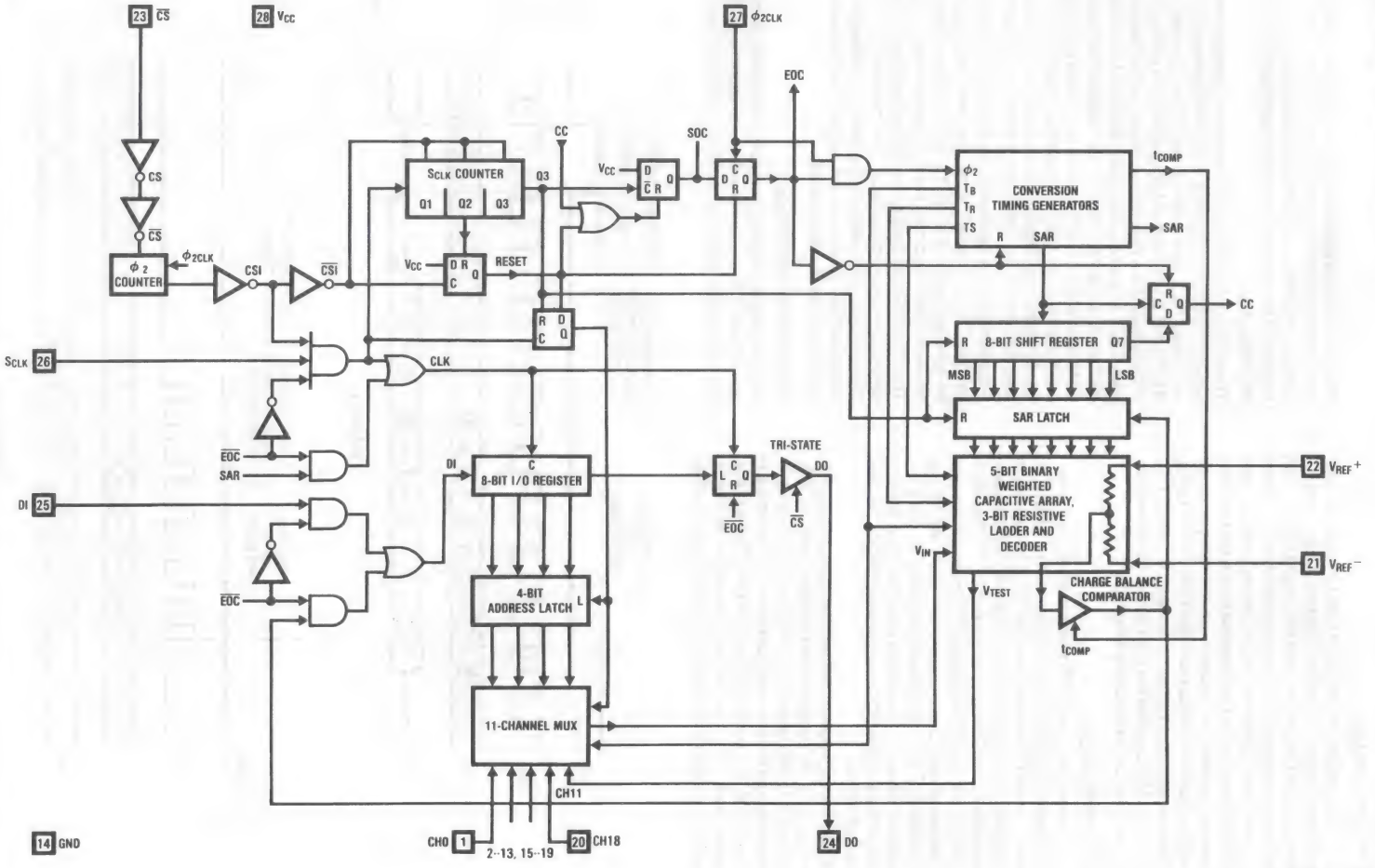
TL/H/9287–14

Channel Addressing Table

TABLE I. ADC 0819 Channel Addressing

MUX ADDRESS								ANALOG CHANNEL SELECTED
A_7	A_6	A_5	A_4	A_3	A_2	A_1	A_0	
0	0	0	0	0	X	X	X	CH0
0	0	0	0	1	X	X	X	CH1
0	0	0	1	0	X	X	X	CH2
0	0	0	1	1	X	X	X	CH3
0	0	1	0	0	X	X	X	CH4
0	0	1	0	1	X	X	X	CH5
0	0	1	1	0	X	X	X	CH6
0	0	1	1	1	X	X	X	CH7
0	1	0	0	0	X	X	X	CH8
0	1	0	0	1	X	X	X	CH9
0	1	0	1	0	X	X	X	CH10
0	1	0	1	1	X	X	X	CH11
0	1	1	0	0	X	X	X	CH12
0	1	1	0	1	X	X	X	CH13
0	1	1	1	0	X	X	X	CH14
0	1	1	1	1	X	X	X	CH15
1	0	0	0	0	X	X	X	CH16
1	0	0	0	1	X	X	X	CH17
1	0	0	1	0	X	X	X	CH18
1	0	0	1	1	X	X	X	V_{TEST}
1	0	1	0	0	X	X	X	No Channel Select
1	0	1	0	1	X	X	X	No Channel Select
1	0	1	1	0	X	X	X	No Channel Select
1	0	1	1	1	X	X	X	No Channel Select
1	1	X	X	X	X	X	X	Logic Test Mode*

*Analog channel inputs CH0 thru CH4 are logic outputs


23 CS
28 VCC
27 φ2CLK
EOC
CC
SOC
SCLK COUNTER
Q1 Q2 Q3
Q3
VCC
CS
CS
φ2 COUNTER
φ2CLK
CSI
CSI
VCC
RESET
φ2
TB
TR
TS
CONVERSION TIMING GENERATORS
R SAR
tCOMP
SAR
CC
R 8-BIT SHIFT REGISTER Q7
MSB
LSB
R SAR LATCH
SCLK 26
CLK
EOC
SAR
DI 25
DI
C
8-BIT I/O REGISTER
TRI-STATE
DO
EOC
CS
5-BIT BINARY WEIGHTED CAPACITIVE ARRAY, 3-BIT RESISTIVE LADDER AND DECODER
22 VREF+
21 VREF−
VIN
VTEST
CHARGE BALANCE COMPARATOR
tCOMP
4-BIT ADDRESS LATCH
11-CHANNEL MUX
CH11
14 GND
CH0 1
2–13, 15–19
20 CH18
24 DO
TL/H/9287–15

Functional Description

1.0 DIGITAL INTERFACE

The ADC0819 uses five input/output pins to implement the serial interface. Taking chip select ($\overline{CS}$) low enables the I/O data lines (DO and DI) and the serial clock input (S_{CLK}). The result of the last conversion is transmitted by the A/D on the DO line, while simultaneously the DI line receives the address data that selects the mux channel for the next conversion. The mux address is shifted in on the rising edge of S_{CLK} and the conversion data is shifted out on the falling edge. It takes eight S_{CLK} cycles to complete the serial I/O. A second clock (ϕ_2) controls the SAR during the conversion process and must be continuously enabled.

1.1 CONTINUOUS S_{CLK}

With a continuous S_{CLK} input $\overline{CS}$ must be used to synchronize the serial data exchange (see *Figure 1*). The ADC0819 recognizes a valid $\overline{CS}$ one to three ϕ_2 clock periods after the actual falling edge of $\overline{CS}$. This is implemented to ensure noise immunity of the $\overline{CS}$ signal. Any spikes on $\overline{CS}$ less than one ϕ_2 clock period will be ignored. $\overline{CS}$ must remain low during the complete I/O exchange which takes eight S_{CLK} cycles. Although $\overline{CS}$ is not immediately acknowledged for the purpose of starting a new conversion, the falling edge of $\overline{CS}$ immediately enables DO to output the MSB (D7) of the previous conversion.

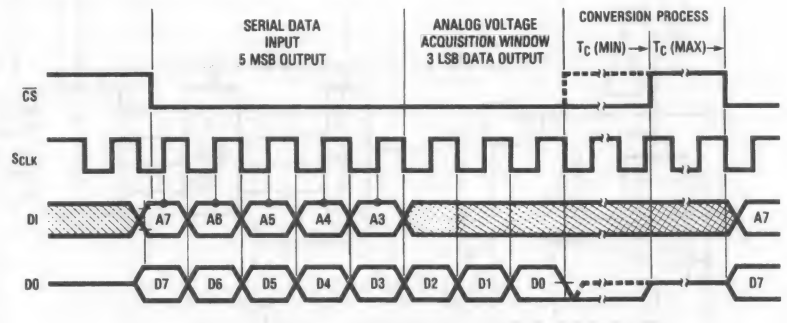
The first S_{CLK} rising edge will be acknowledged after a set-up time ($t_{set\text{-}up}$) has elapsed from the falling edge of $\overline{CS}$. This and the following seven S_{CLK} rising edges will shift in the channel address for the analog multiplexer. Since there are 19 channels only five address bits are utilized. The first five S_{CLK} cycles clock in the mux address, during the next three S_{CLK} cycles the analog input is selected and sampled. During this mux address/sample cycle, data from the last conversion is also clocked out on DO. Since D7 was clocked out on the falling edge of $\overline{CS}$ only data bits D6–D0 remain to be received. The following seven falling edges of S_{CLK} shift out this data on DO.

The 8th S_{CLK} falling edge initiates the beginning of the A/D's actual conversion process which takes between 26 and 32 ϕ_2 cycles (T_C). During this time $\overline{CS}$ can go high to TRI-STATE DO and disable the S_{CLK} input or it can remain low. If $\overline{CS}$ is held low a new I/O exchange will not start until the conversion sequence has been completed, however once the conversion ends serial I/O will immediately begin. Since there is an ambiguity in the conversion time (T_C) synchronizing the data exchange is impossible. Therefore $\overline{CS}$ should go high before the 26th ϕ_2 clock has elasped and return low after the 32nd ϕ_2 to synchronize serial communication.

A conversion or I/O operation can be aborted at any time by strobing $\overline{CS}$. If $\overline{CS}$ is high or low less than one ϕ_2 clock it will be ignored by the A/D. If the $\overline{CS}$ is strobed high or low between 1 to 3 ϕ_2 clocks the A/D may or may not respond. Therefore $\overline{CS}$ must be strobed high or low greater than 3 ϕ_2 clocks to ensure recognition. If a conversion or I/O exchange is aborted while in process the consequent data output will be erroneous until a complete conversion sequence has been implemented.

1.2 DISCONTINUOUS S_{CLK}

Another way to accomplish synchronous serial communication is to tie $\overline{CS}$ low continuously and disable S_{CLK} after its 8th falling edge (see *Figure 2*). S_{CLK} must remain low for



TL/H/9287–16

FIGURE 1

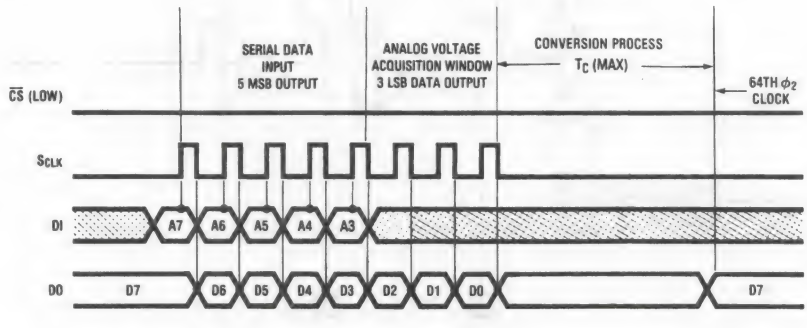


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FIGURE 2

Functional Description (Continued)

at least 32 ϕ_2 clocks to ensure that the A/D has completed its conversion. If S_{CLK} is enabled sooner, synchronizing to the data output on DO is not possible since an end of conversion signal from the A/D is not available and the actual conversion time is not known. With $\overline{CS}$ low during the conversion time (32 ϕ_2 max) DO will go high or low after the eighth falling edge of S_{CLK} until the conversion is completed. Once the conversion is through DO will transmit the MSB. The rest of the data will be shifted out once S_{CLK} is enabled as discussed previously.

If $\overline{CS}$ goes high during the conversion sequence DO is tristated, and the result is not affected so long as $\overline{CS}$ remains high until the end of the conversion.

1.2 MULTIPLEXER ADDRESSING

The five bit mux address is shifted, MSB first, into DI. Input data corresponds to the channel selected as shown in table 1. Care should be taken not to send an address greater than or equal to twenty four (11XXX) as this puts the A/D in a digital testing mode. In this mode the analog inputs CH0 thru CH4 become digital outputs, for our use in production testing.

2.0 ANALOG INPUT

2.1 THE INPUT SAMPLE AND HOLD

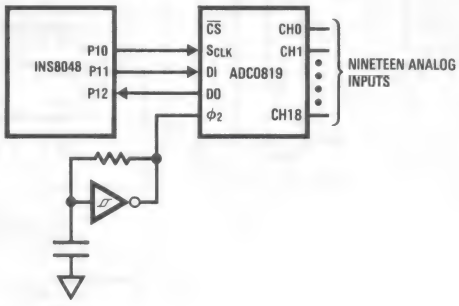
The ADC0819's sample/hold capacitor is implemented in its capacitive ladder structure. After the channel address is received, the ladder is switched to sample the proper analog input. This sampling mode is maintained for 1 μsec after the eighth S_{CLK} falling edge. The hold mode is initiated with the start of the conversion process. An acquisition window of $3t_{SCLK}+1$ μsec is therefore available to allow the ladder capacitance to settle to the analog input voltage. Any change in the analog voltage before or after the acquisition window will not effect the A/D conversion result.

In the most simple case, the ladder's acquisition time is determined by the R_{on} (3K) of the multiplexer switches and the total ladder capacitance (90pf). These values yield an acquisition time of about 2 μsec for a full scale reading. Therefore the analog input must be stable for at least 2 μsec before and 1 μsec after the eighth S_{CLK} falling edge to ensure a proper conversion. External input source resistance and capacitance will lengthen the acquisition time and should be accounted for.

Other conventional sample and hold error specifications are included in the error and timing specs of the A/D. The hold step and gain error sample/hold specs are taken into account in the ADC0819's total unadjusted error, while the hold settling time is included in the A/D's max conversion time of 32 ϕ_2 clock periods. The hold droop rate can be thought of as being zero since an unlimited amount of time can pass between a conversion and the reading of data. However, once the data is read it is lost and another conversion is started.

Typical Applications

ADC0819-INS8048 INTERFACE



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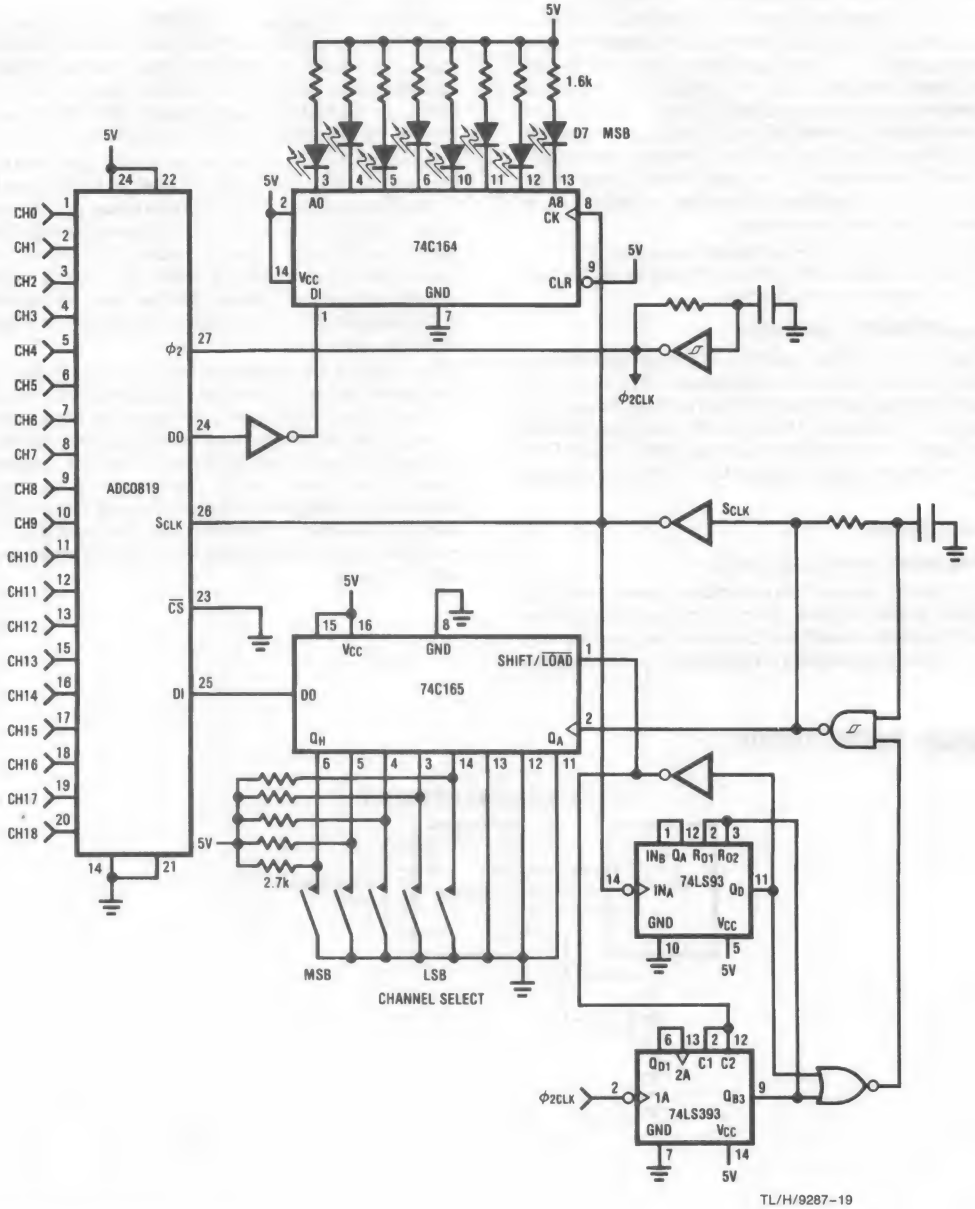
ADC0819 FUNCTIONAL CIRCUIT



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Ordering Information

Temperature Range		0°C to +70°C	−40°C to +85°C
Total Unadjusted Error	±½ LSB	ADC0819BCN	ADC0819BCV
	±1 LSB	ADC0819CCN	ADC0819CCV
Package Outline		N28B	V28A

National Semiconductor

microCMOS

ADC0820 8-Bit High Speed μP Compatible A/D Converter with Track/Hold Function

General Description

By using a half-flash conversion technique, the 8-bit ADC0820 CMOS A/D offers a 1.5 μs conversion time and dissipates only 75 mW of power. The half-flash technique consists of 32 comparators, a most significant 4-bit ADC and a least significant 4-bit ADC.

The input to the ADC0820 is tracked and held by the input sampling circuitry eliminating the need for an external sample-and-hold for signals moving at less than 100 mV/μs.

For ease of interface to microprocessors, the ADC0820 has been designed to appear as a memory location or I/O port without the need for external interfacing logic.

Key Specifications

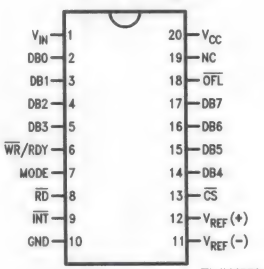
- Resolution — 8 Bits
- Conversion Time — 2.5 μs Max (RD Mode); 1.5 μs Max (WR-RD Mode)
- Input signals with slew rate of 100 mV/μs converted without external sample-and-hold to 8 bits
- Low Power — 75 mW Max
- Total Unadjusted Error — ± ½ LSB and ± 1 LSB

Features

- Built-in track-and-hold function
- No missing codes
- No external clocking
- Single supply—5 V_{DC}
- Easy interface to all microprocessors, or operates stand-alone
- Latched TRI-STATE® output
- Logic inputs and outputs meet both MOS and T^2L voltage level specifications
- Operates ratiometrically or with any reference value equal to or less than V_{CC}
- 0V to 5V analog input voltage range with single 5V supply
- No zero or full-scale adjust required
- Overflow output available for cascading
- 0.3″ standard width 20-pin DIP
- 20-pin molded chip carrier package
- 20-pin small outline package

Connection and Functional Diagrams

Dual-In-Line and Small Outline Packages



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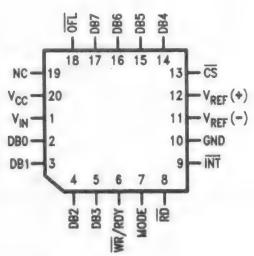
Top View

Molded Chip Carrier Package



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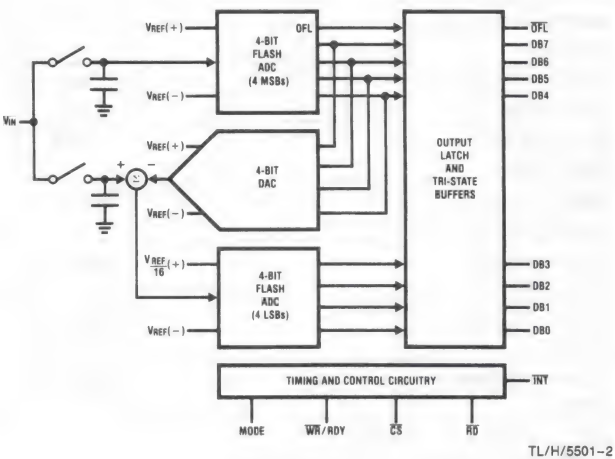

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FIGURE 1

See Ordering Information

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	10V
Logic Control Inputs	$-0.2V$ to $V_{CC}+0.2V$
Voltage at Other Inputs and Output	$-0.2V$ to $V_{CC}+0.2V$
Storage Temperature Range	$-65°C$ to $+150°C$
Package Dissipation at $T_A = 25°C$	875 mW
Input Current at Any Pin (Note 5)	1 mA
Package Input Current (Note 5)	4 mA
ESD Susceptability (Note 9)	1200V
Lead Temp. (Soldering, 10 sec.)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
Surface Mount Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

Operating Ratings (Notes 1 & 2)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0820BD, ADC0820CJ	$-55°C \le T_A \le +125°C$
ADC0820BCD, ADC0820CCJ	$-40°C \le T_A \le +85°C$
ADC0820BCN, ADC0820CCN	$0°C \le T_A \le 70°C$
ADC0820BCV, ADC0820CCV	$0°C \le T_A \le 70°C$
ADC0820BCWM, ADC0820CCWM	$0°C \le T_A \le 70°C$
V_{CC} Range	4.5V to 8V

Converter Characteristics

The following specifications apply for RD mode (pin 7 = 0), $V_{CC} = 5V$, $V_{REF}(+) = 5V$, and $V_{REF}(-) = GND$ unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX}**; all other limits $T_A = T_j = 25°C$.

Parameter	Conditions	ADC0820BD, ADC0820CJ ADC0820BCD, ADC0820CCJ			ADC0820BCN, ADC0820CCN ADC0820BCV, ADC0820CCV ADC0820BCWM, ADC0820CCWM			Limit Units
		Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	
Resolution			**8**			8	**8**	Bits
Total Unadjusted Error (Note 3)	ADC0820BD, BCD ADC0820BCN ADC0820CD, CCD ADC0820CCN		**±½** **±1**			 ±½ ±1	 **±½** **±1**	LSB LSB LSB LSB
Minimum Reference Resistance		2.3	**1.00**		2.3	1.2		kΩ
Maximum Reference Resistance		2.3	**6**		2.3	5.3	**6**	kΩ
Maximum $V_{REF}(+)$ Input Voltage			**V_{CC}**			V_{CC}	**V_{CC}**	V
Minimum $V_{REF}(-)$ Input Voltage			**GND**			GND	**GND**	V
Minimum $V_{REF}(+)$ Input Voltage			**$V_{REF}(-)$**			$V_{REF}(-)$	**$V_{REF}(-)$**	V
Maximum $V_{REF}(-)$ Input Voltage			**$V_{REF}(+)$**			$V_{REF}(+)$	**$V_{REF}(+)$**	V
Maximum V_{IN} Input Voltage			**$V_{CC}+0.1$**			$V_{CC}+0.1$	**$V_{CC}+0.1$**	V
Minimum V_{IN} Input Voltage			**GND−0.1**			GND−0.1	**GND−0.1**	V
Maximum Analog Input Leakage Current	$\overline{CS} = V_{CC}$ $V_{IN} = V_{CC}$ $V_{IN} = GND$		 **3** **−3**			 0.3 −0.3	 **3** **−3**	 μA μA
Power Supply Sensitivity	$V_{CC} = 5V \pm 5\%$	±1/16	±¼		±1/16	±¼	±¼	LSB

DC Electrical Characteristics

The following specifications apply for $V_{CC}=5V$, unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX}**; all other limits $T_A=T_J=25°C$.

Parameter	Conditions		ADC0820BD, ADC0820CJ ADC0820BCD, ADC0820CCJ			ADC0820BCN, ADC0820CCN ADC0820BCV, ADC0820CCV ADC0820BCWM, ADC0820CCWM			Limit Units
			Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	
$V_{IN(1)}$, Logical "1" Input Voltage	$V_{CC}=5.25V$	$\overline{CS}$, $\overline{WR}$, $\overline{RD}$		**2.0**			2.0	**2.0**	V
		Mode		**3.5**			3.5	**3.5**	V
$V_{IN(0)}$, Logical "0" Input Voltage	$V_{CC}=4.75V$	$\overline{CS}$, $\overline{WR}$, $\overline{RD}$		**0.8**			0.8	**0.8**	V
		Mode		**1.5**			1.5	**1.5**	V
$I_{IN(1)}$, Logical "1" Input Current	$V_{IN(1)}=5V$; $\overline{CS}$, $\overline{RD}$		0.005	**1**		0.005		**1**	μA
	$V_{IN(1)}=5V$; $\overline{WR}$		0.1	**3**		0.1	0.3	**3**	μA
	$V_{IN(1)}=5V$; Mode		50	**200**		50	170	**200**	μA
$I_{IN(0)}$, Logical "0" Input Current	$V_{IN(0)}=0V$; $\overline{CS}$, $\overline{RD}$, $\overline{WR}$, Mode		−0.005	**−1**		−0.005		**−1**	μA
$V_{OUT(1)}$, Logical "1" Output Voltage	$V_{CC}=4.75V$, $I_{OUT}=-360$ μA; DB0–DB7, $\overline{OFL}$, $\overline{INT}$			**2.4**			2.8	**2.4**	V
	$V_{CC}=4.75V$, $I_{OUT}=-10$ μA; DB0–DB7, $\overline{OFL}$, $\overline{INT}$			**4.5**			4.6	**4.5**	V
$V_{OUT(0)}$, Logical "0" Output Voltage	$V_{CC}=4.75V$, $I_{OUT}=1.6$ mA; DB0–DB7, $\overline{OFL}$, $\overline{INT}$, RDY			**0.4**			0.34	**0.4**	V
I_{OUT}, TRI-STATE Output Current	$V_{OUT}=5V$; DB0–DB7, RDY		0.1	**3**		0.1	0.3	**3**	μA
	$V_{OUT}=0V$; DB0–DB7, RDY		−0.1	**−3**		−0.1	−0.3	**−3**	μA
I_{SOURCE}, Output Source Current	$V_{OUT}=0V$; DB0–DB7, $\overline{OFL}$ $\overline{INT}$		−12	**−6**		−12	−7.2	**−6**	mA
			−9	**−4.0**		−9	−5.3	**−4.0**	mA
I_{SINK}, Output Sink Current	$V_{OUT}=5V$; DB0–DB7, $\overline{OFL}$, $\overline{INT}$, RDY		14	**7**		14	8.4	**7**	mA
I_{CC}, Supply Current	$\overline{CS}=\overline{WR}=\overline{RD}=0$		7.5	**15**		7.5	13	**15**	mA

AC Electrical Characteristics

The following specifications apply for $V_{CC}=5V$, $t_r=t_f=20$ ns, $V_{REF}(+)=5V$, $V_{REF}(-)=0V$ and $T_A=25°C$ unless otherwise specified.

Parameter		Conditions	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
t_{CRD}, Conversion Time for RD Mode		Pin 7 = 0, *(Figure 2)*	1.6		2.5	μs
t_{ACC0}, Access Time (Delay from Falling Edge of $\overline{RD}$ to Output Valid)		Pin 7 = 0, *(Figure 2)*	$t_{CRD}+20$		$t_{CRD}+50$	ns
t_{CWR-RD}, Conversion Time for WR-RD Mode		Pin 7 = V_{CC}; t_{WR} = 600 ns, t_{RD} = 600 ns; *(Figures 3a* and *3b)*			1.52	μs
t_{WR}, Write Time	Min	Pin 7 = V_{CC}; *(Figures 3a* and *3b)*		600		ns
	Max	(Note 4) See Graph	50			μs
t_{RD}, Read Time	Min	Pin 7 = V_{CC}; *(Figures 3a* and *3b)* (Note 4) See Graph		600		ns
t_{ACC1}, Access Time (Delay from Falling Edge of $\overline{RD}$ to Output Valid)		Pin 7 = V_{CC}, $t_{RD}<t_l$; *(Figure 3a)*				
		C_L = 15 pF	190		280	ns
		C_L = 100 pF	210		320	ns
t_{ACC2}, Access Time (Delay from Falling Edge of $\overline{RD}$ to Output Valid)		Pin 7 = V_{CC}, $t_{RD}>t_l$; *(Figure 3b)*				
		C_L = 15 pF	70		120	ns
		C_L = 100 pF	90		150	ns

AC Electrical Characteristics (Continued) The following specifications apply for $V_{CC}=5V$, $t_r=t_f=20$ ns, $V_{REF}(+)=5V$, $V_{REF}(-)=0V$ and $T_A=25°C$ unless otherwise specified.

Parameter	Conditions	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
t_i, Internal Comparison Time	Pin 7 = V_{CC}; *(Figures 3b* and *4)* $C_L=50$ pF	800		1300	ns
t_{1H}, t_{0H}, TRI-STATE Control (Delay from Rising Edge of $\overline{RD}$ to Hi-Z State)	$R_L=1k$, $C_L=10$ pF	100		200	ns
$t_{\overline{INT}L}$, Delay from Rising Edge of $\overline{WR}$ to Falling Edge of $\overline{INT}$	Pin 7 = V_{CC}, $C_L=50$ pF $t_{RD}>t_i$; *(Figure 3b)* $t_{RD}<t_i$; *(Figure 3a)*	 $t_{RD}+200$		 t_i $t_{RD}+290$	 ns ns
$t_{\overline{INT}H}$, Delay from Rising Edge of $\overline{RD}$ to Rising Edge of $\overline{INT}$	*(Figures 2, 3a* and *3b)* $C_L=50$ pF	125		225	ns
$t_{\overline{INT}HWR}$, Delay from Rising Edge of $\overline{WR}$ to Rising Edge of $\overline{INT}$	*(Figure 4)*, $C_L=50$ pF	175		270	ns
t_{RDY}, Delay from $\overline{CS}$ to RDY	*(Figure 2)*, $C_L=50$ pF, Pin 7 = 0	50		100	ns
t_{ID}, Delay from $\overline{INT}$ to Output Valid	*(Figure 4)*	20		50	ns
t_{RI}, Delay from $\overline{RD}$ to $\overline{INT}$	Pin 7 = V_{CC}, $t_{RD}<t_i$ *(Figure 3a)*	200		290	ns
t_P, Delay from End of Conversion to Next Conversion	*(Figures 2, 3a, 3b* and *4)* (Note 4) See Graph			500	ns
Slew Rate, Tracking		0.1			V/μs
C_{VIN}, Analog Input Capacitance		45			pF
C_{OUT}, Logic Output Capacitance		5			pF
C_{IN}, Logic Input Capacitance		5			pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to the GND pin, unless otherwise specified.

Note 3: Total unadjusted error includes offset, full-scale, and linearity errors.

Note 4: Accuracy may degrade if t_{WR} or t_{RD} is shorter than the minimum value specified. See Accuracy vs t_{WR} and Accuracy vs t_{RD} graphs.

Note 5: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN}<V^-$ or $V_{IN}>V^+$) the absolute value of current at that pin should be limited to 1 mA or less. The 4 mA package input current limits the number of pins that can exceed the power supply boundaries with a 1 mA current limit to four.

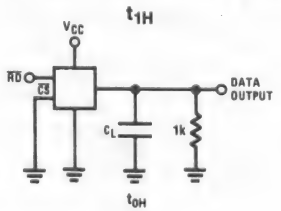
Note 6: Typicals are at 25°C and represent most likely parametric norm.

Note 7: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 8: Design limits are guaranteed but not 100% tested. These limits are not used to calculate outgoing quality levels.

Note 9: Human body model, 100 pF discharaged through a 1.5 kΩ resistor.

TRI-STATE Test Circuits and Waveforms



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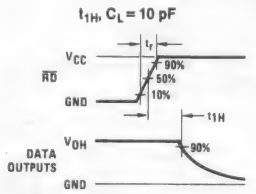

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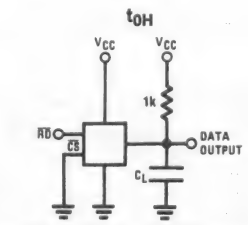

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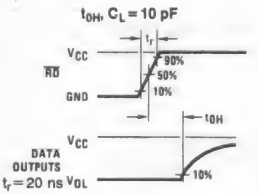
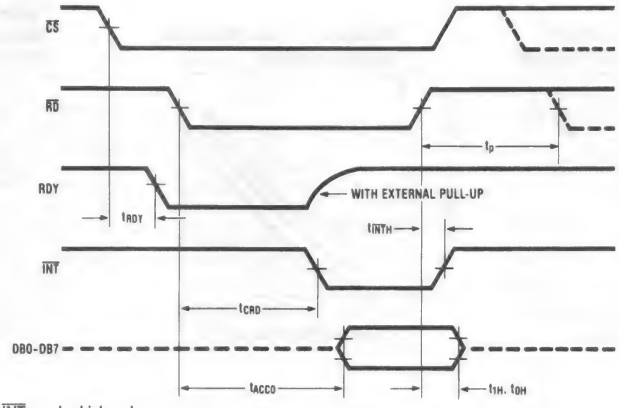


TL/H/5501–6

Timing Diagrams



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Note: On power-up the state of $\overline{INT}$ can be high or low.

FIGURE 2. RD Mode (Pin 7 is Low)

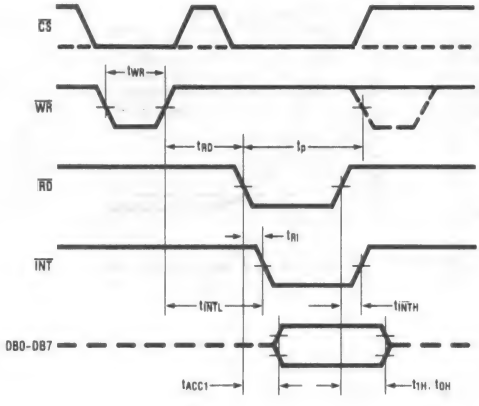

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FIGURE 3a. WR-RD Mode (Pin 7 is High and $t_{RD} < t_I$)

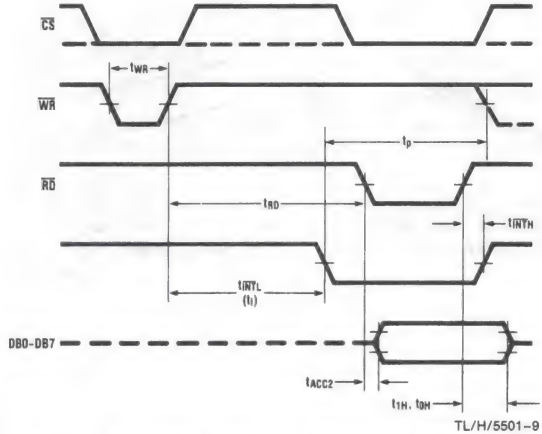

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FIGURE 3b. WR-RD Mode (Pin 7 is High and $t_{RD} > t_I$)

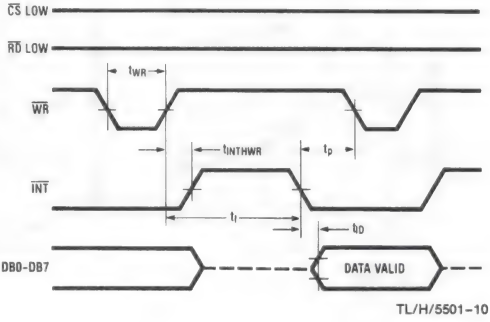

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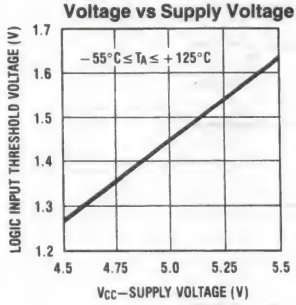
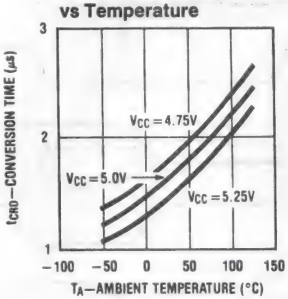
FIGURE 4. WR-RD Mode (Pin 7 is High) Stand-Alone Operation

Typical Performance Characteristics

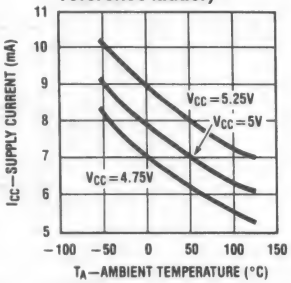
Logic Input Threshold Voltage vs Supply Voltage



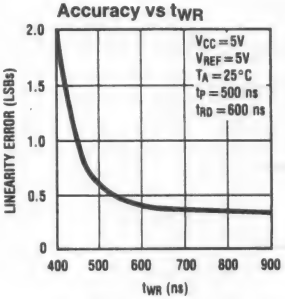
Conversion Time (RD Mode) vs Temperature



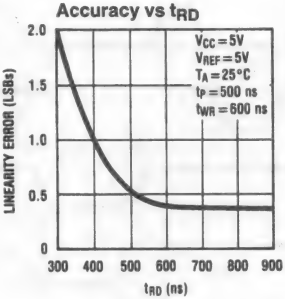
Power Supply Current vs Temperature (not including reference ladder)



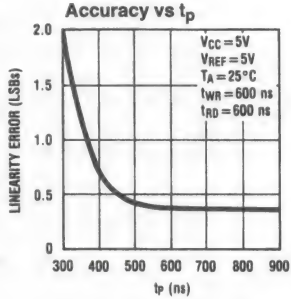
Accuracy vs t_{WR}



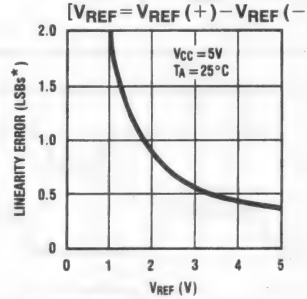
Accuracy vs t_{RD}



Accuracy vs t_P

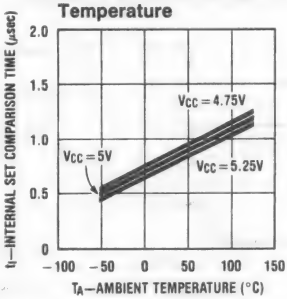


Accuracy vs V_{REF}
$[V_{REF} = V_{REF}(+) - V_{REF}(-)]$

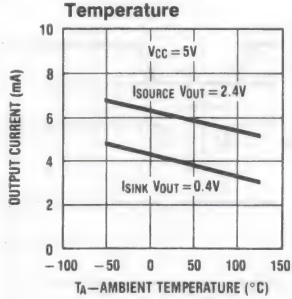


*1 LSB = $\frac{V_{REF}}{256}$

t_I, Internal Time Delay vs Temperature



Output Current vs Temperature



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Description of Pin Functions

Pin	Name	Function
1	V_{IN}	Analog input; range = $GND \leq V_{IN} \leq V_{CC}$
2	DB0	TRI-STATE data output—bit 0 (LSB)
3	DB1	TRI-STATE data output—bit 1
4	DB2	TRI-STATE data output—bit 2
5	DB3	TRI-STATE data output—bit 3
6	$\overline{WR}$/RDY	**WR-RD Mode** **$\overline{WR}$:** With $\overline{CS}$ low, the conversion is started on the falling edge of $\overline{WR}$. Approximately 800 ns (the preset internal time out, t_I) after the $\overline{WR}$ rising edge, the result of the conversion will be strobed into the output latch, provided that $\overline{RD}$ does not occur prior to this time out (see *Figures 3a* and *3b*). **RD Mode** **RDY:** This is an open drain output (no internal pull-up device). RDY will go low after the falling edge of $\overline{CS}$; RDY will go TRI-STATE when the result of the conversion is strobed into the output latch. It is used to simplify the interface to a microprocessor system (see *Figure 2*).
7	Mode	**Mode:** Mode selection input—it is internally tied to GND through a 50 μA current source. **RD Mode:** When mode is low **WR-RD Mode:** When mode is high
8	$\overline{RD}$	**WR-RD Mode** With $\overline{CS}$ low, the TRI-STATE data outputs (DB0-DB7) will be activated when $\overline{RD}$ goes low (see *Figure 4*). $\overline{RD}$ can also be used to increase the speed of the converter by reading data prior to the preset internal time out (t_I, ~800 ns). If this is done, the data result transferred to output latch is latched after the falling edge of the $\overline{RD}$ (see *Figures 3a* and *3b*). **RD Mode** With $\overline{CS}$ low, the conversion will start with $\overline{RD}$ going low, also $\overline{RD}$ will enable the TRI-STATE data outputs at the completion of the conversion. RDY going TRI-STATE and $\overline{INT}$ going low indicates the completion of the conversion (see *Figure 2*).
9	$\overline{INT}$	**WR-RD Mode** $\overline{INT}$ going low indicates that the conversion is completed and the data result is in the output latch. $\overline{INT}$ will go low, ~800 ns (the preset internal time out, t_I) after the rising edge of $\overline{WR}$ (see *Figure 3b*); or $\overline{INT}$ will go low after the falling edge of $\overline{RD}$, if $\overline{RD}$ goes low prior to the 800 ns time out (see *Figure 3a*). $\overline{INT}$ is reset by the rising edge of $\overline{RD}$ or $\overline{CS}$ (see *Figures 3a* and *3b*). **RD Mode** $\overline{INT}$ going low indicates that the conversion is completed and the data result is in the output latch. $\overline{INT}$ is reset by the rising edge of $\overline{RD}$ or $\overline{CS}$ (see *Figure 2*).
10	GND	Ground
11	$V_{REF}(-)$	The bottom of resistor ladder, voltage range: $GND \leq V_{REF}(-) \leq V_{REF}(+)$ (Note 5)
12	$V_{REF}(+)$	The top of resistor ladder, voltage range: $V_{REF}(-) \leq V_{REF}(+) \leq V_{CC}$ (Note 5)
13	$\overline{CS}$	$\overline{CS}$ must be low in order for the $\overline{RD}$ or $\overline{WR}$ to be recognized by the converter.
14	DB4	TRI-STATE data output—bit 4
15	DB5	TRI-STATE data output—bit 5
16	DB6	TRI-STATE data output—bit 6
17	DB7	TRI-STATE data output—bit 7 (MSB)
18	$\overline{OFL}$	Overflow output—If the analog input is higher than the $V_{REF}(+)$, $\overline{OFL}$ will be low at the end of conversion. It can be used to cascade 2 or more devices to have more resolution (9, 10-bit). This output is always active and does not go into TRI-STATE as DB0–DB7 do.
19	NC	No connection
20	V_{CC}	Power supply voltage

1.0 Functional Description

1.1 GENERAL OPERATION

The ADC0820 uses two 4-bit flash A/D converters to make an 8-bit measurement (*Figure 1*). Each flash ADC is made up of 15 comparators which compare the unknown input to a reference ladder to get a 4-bit result. To take a full 8-bit reading, one flash conversion is done to provide the 4 most significant data bits (via the MS flash ADC). Driven by the 4 MSBs, an internal DAC recreates an analog approximation of the input voltage. This analog signal is then subtracted from the input, and the difference voltage is converted by a second 4-bit flash ADC (the LS ADC), providing the 4 least significant bits of the output data word.

The internal DAC is actually a subsection of the MS flash converter. This is accomplished by using the same resistor ladder for the A/D as well as for generating the DAC signal. The DAC output is actually the tap on the resistor ladder which most closely approximates the analog input. In addition, the "sampled-data" comparators used in the ADC0820 provide the ability to compare the magnitudes of several analog signals simultaneously, without using input summing amplifiers. This is especially useful in the LS flash ADC, where the signal to be converted is an analog difference.

1.0 Functional Description (Continued)

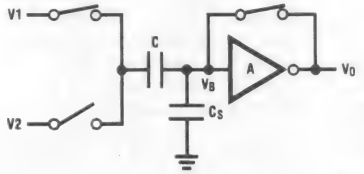
1.2 THE SAMPLED-DATA COMPARATOR

Each comparator in the ADC0820 consists of a CMOS inverter with a capacitively coupled input (*Figure 5*). Analog switches connect the two comparator inputs to the input capacitor (C) and also connect the inverter's input and output. This device in effect now has one differential input pair. A comparison requires two cycles, one for zeroing the comparator, and another for making the comparison.

In the first cycle, one input switch and the inverter's feedback switch (*Figure 5a*) are closed. In this interval, C is charged to the connected input (V1) less the inverter's bias voltage (V_B, approximately 1.2V). In the second cycle (*Figure 5b*), these two switches are opened and the other (V2) input's switch is closed. The input capacitor now subtracts its stored voltage from the second input and the difference is amplified by the inverter's open loop gain. The inverter's input (V_B') becomes

$$V_B - (V1 - V2)\frac{C}{C + C_S}$$

and the output will go high or low depending on the sign of $V_B' - V_B$.



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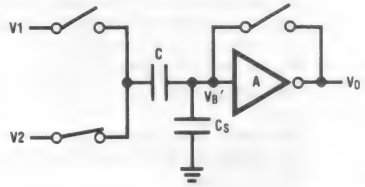
- $V_O = V_B$
- V on C = $V1 - V_B$
- C_S = stray input node capacitor
- V_B = inverter input bias voltage

FIGURE 5a. Zeroing Phase



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- $V_B' - V_B = (V2 - V1)\frac{C}{C + C_S}$
- $V_O' = \frac{-A}{C + C_S}[CV2 - CV1]$
- V_O' is dependent on V2 − V1

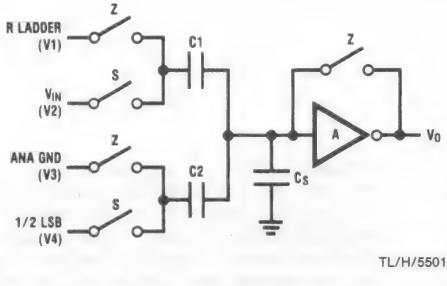
FIGURE 5b. Compare Phase

FIGURE 5. Sampled-Data Comparator

The actual circuitry used in the ADC0820 is a simple but important expansion of the basic comparator described above. By adding a second capacitor and another set of switches to the input (*Figure 6*), the scheme can be expanded to make dual differential comparisons. In this circuit, the feedback switch and one input switch on each capacitor (Z switches) are closed in the zeroing cycle. A comparison is then made by connecting the second input on each capacitor and opening all of the other switches (S switches). The change in voltage at the inverter's input, as a result of the change in charge on each input capacitor, will now depend on both input signal differences.

1.3 ARCHITECTURE

In the ADC0820, one bank of 15 comparators is used in each 4-bit flash A/D converter (*Figure 7*). The MS (most significant) flash ADC also has one additional comparator to detect input overrange. These two sets of comparators operate alternately, with one group in its zeroing cycle while the other is comparing.



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$$V_O = \frac{-A}{C1 + C2 + C_S}[C1(V2 - V1) + C2(V4 - V3)]$$

$$= \frac{-A}{C1 + C2 + C_S}[\Delta Q_{C1} + \Delta Q_{C2}]$$

FIGURE 6. ADC0820 Comparator (from MS Flash ADC)

Detailed Block Diagram

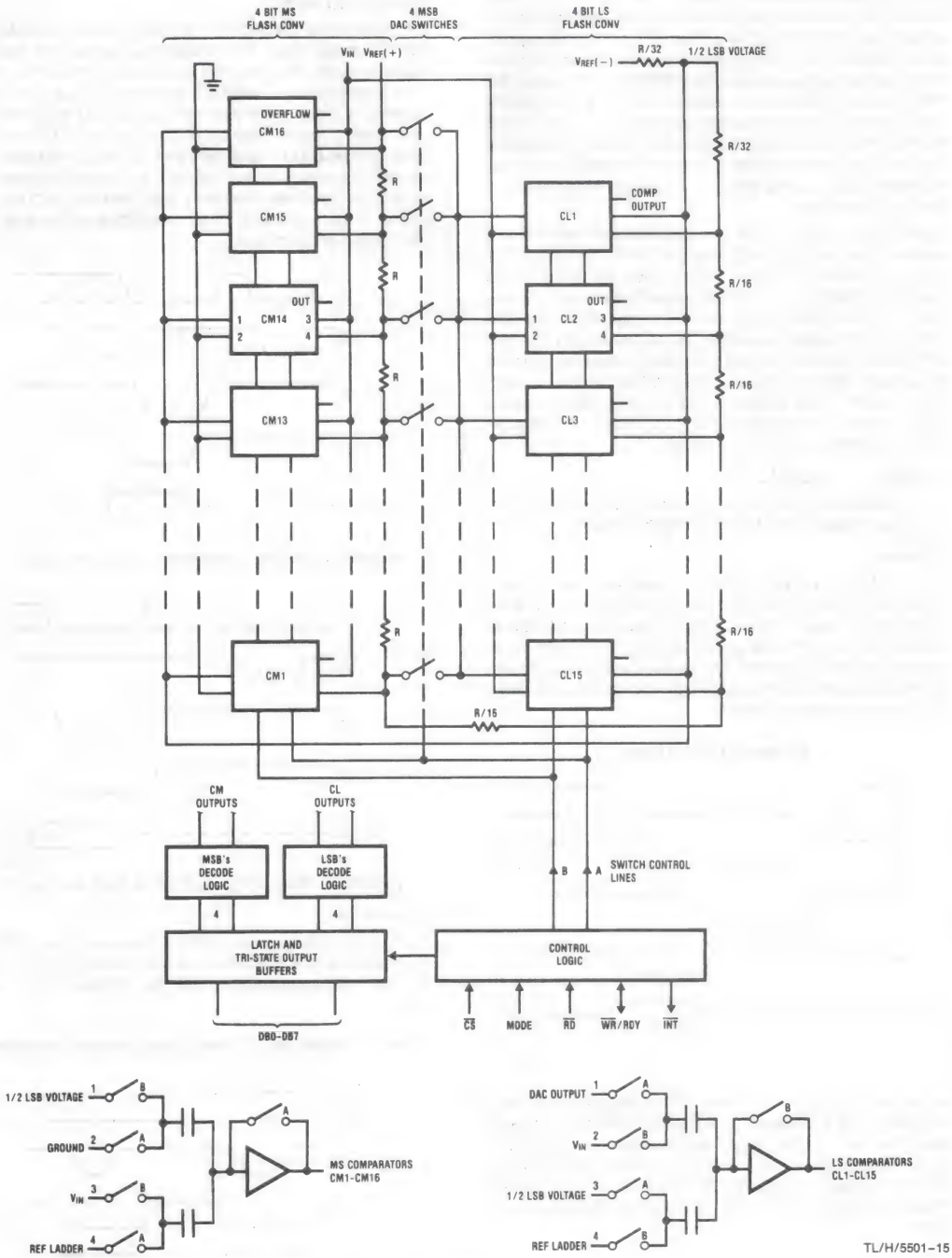


FIGURE 7

1.0 Functional Description (Continued)

When a typical conversion is started, the $\overline{WR}$ line is brought low. At this instant the MS comparators go from zeroing to comparison mode (*Figure 8*). When $\overline{WR}$ is returned high after at least 600 ns, the output from the first set of comparators (the first flash) is decoded and latched. At this point the two 4-bit converters change modes and the LS (least significant) flash ADC enters its compare cycle. No less than 600 ns later, the $\overline{RD}$ line may be pulled low to latch the lower 4 data bits and finish the 8-bit conversion. When $\overline{RD}$ goes low, the flash A/Ds change state once again in preparation for the next conversion.

Figure 8 also outlines how the converter's interface timing relates to its analog input (V_{IN}). In WR-RD mode, V_{IN} is measured while $\overline{WR}$ is low. In RD mode, sampling occurs during the first 800 ns of $\overline{RD}$. Because of the input connections to the ADC0820's LS and MS comparators, the converter has the ability to sample V_{IN} at one instant (Section 2.4), despite the fact that two separate 4-bit conversions are being done. More specifically, when $\overline{WR}$ is low the MS flash is in compare mode (connected to V_{IN}), and the LS flash is in zero mode (also connected to V_{IN}). Therefore both flash ADCs sample V_{IN} at the same time.

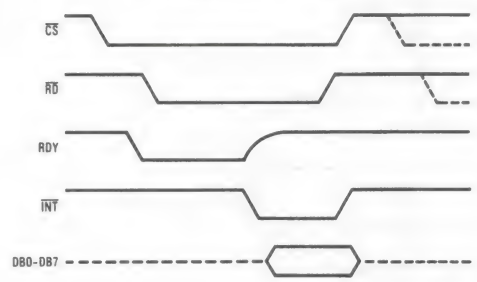
1.4 DIGITAL INTERFACE

The ADC0820 has two basic interface modes which are selected by strapping the MODE pin high or low.

RD Mode

With the MODE pin grounded, the converter is set to Read mode. In this configuration, a complete conversion is done by pulling $\overline{RD}$ low until output data appears. An $\overline{INT}$ line is provided which goes low at the end of the conversion as well as a RDY output which can be used to signal a processor that the converter is busy or can also serve as a system Transfer Acknowledge signal.

RD Mode (Pin 7 is Low)



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When in RD mode, the comparator phases are internally triggered. At the falling edge of $\overline{RD}$, the MS flash converter goes from zero to compare mode and the LS ADC's comparators enter their zero cycle. After 800 ns, data from the MS flash is latched and the LS flash ADC enters compare mode. Following another 800 ns, the lower 4 bits are recovered.

WR then RD Mode

With the MODE pin tied high, the A/D will be set up for the WR-RD mode. Here, a conversion is started with the $\overline{WR}$ input; however, there are two options for reading the output data which relate to interface timing. If an interrupt driven scheme is desired, the user can wait for $\overline{INT}$ to go low before reading the conversion result (*Figure B*). $\overline{INT}$ will typically go low 800 ns after $\overline{WR}$'s rising edge. However, if a shorter conversion time is desired, the processor need not wait for $\overline{INT}$ and can exercise a read after only 600 ns (*Figure A*). If this is done, $\overline{INT}$ will immediately go low and data will appear at the outputs.

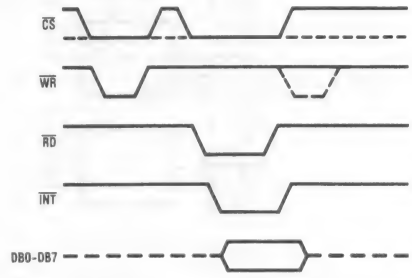

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FIGURE A. WR-RD Mode (Pin 7 is High and $t_{RD} < t_I$)

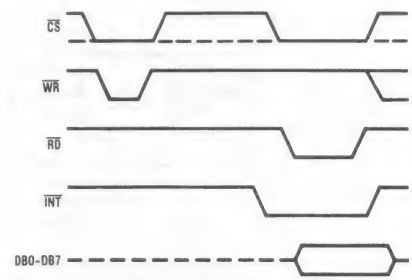

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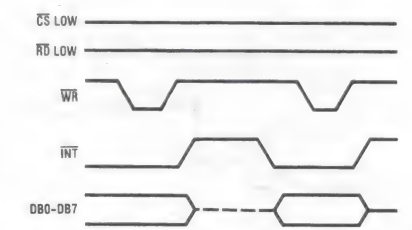
FIGURE B. WR-RD Mode (Pin 7 is High and $t_{RD} > t_I$)

Stand-Alone

For stand-alone operation in WR-RD mode, $\overline{CS}$ and $\overline{RD}$ can be tied low and a conversion can be started with $\overline{WR}$. Data will be valid approximately 800 ns following $\overline{WR}$'s rising edge.

WR-RD Mode (Pin 7 is High) Stand-Alone Operation



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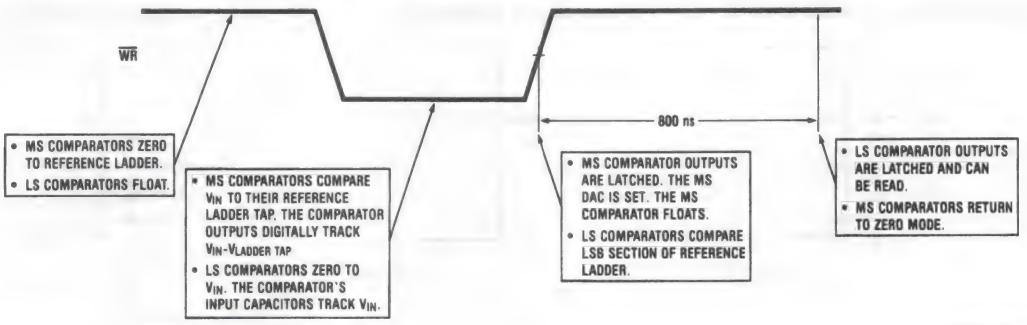
1.0 Functional Description (Continued)


WR
• MS COMPARATORS ZERO TO REFERENCE LADDER.
• LS COMPARATORS FLOAT.
• MS COMPARATORS COMPARE VIN TO THEIR REFERENCE LADDER TAP. THE COMPARATOR OUTPUTS DIGITALLY TRACK VIN-VLADDER TAP
• LS COMPARATORS ZERO TO VIN. THE COMPARATOR'S INPUT CAPACITORS TRACK VIN.
800 ns
• MS COMPARATOR OUTPUTS ARE LATCHED. THE MS DAC IS SET. THE MS COMPARATOR FLOATS.
• LS COMPARATORS COMPARE LSB SECTION OF REFERENCE LADDER.
• LS COMPARATOR OUTPUTS ARE LATCHED AND CAN BE READ.
• MS COMPARATORS RETURN TO ZERO MODE.

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Note: MS means most significant
LS means least significant

FIGURE 8. Operating Sequence (WR-RD Mode)

OTHER INTERFACE CONSIDERATIONS

In order to maintain conversion accuracy, $\overline{WR}$ has a maximum width spec of 50 μs. When the MS flash ADC's sampled-data comparators (Section 1.2) are in comparison mode ($\overline{WR}$ is low), the input capacitors (C, *Figure 6*) must hold their charge. Switch leakage and inverter bias current can cause errors if the comparator is left in this phase for too long.

Since the MS flash ADC enters its zeroing phase at the end of a conversion (Section 1.3), a new conversion cannot be started until this phase is complete. The minimum spec for this time (t_P, *Figures 2, 3a, 3b,* and *4*) is 500 ns.

2.0 Analog Considerations

2.1 REFERENCE AND INPUT

The two V_{REF} inputs of the ADC0820 are fully differential and define the zero to full-scale input range of the A to D converter. This allows the designer to easily vary the span of the analog input since this range will be equivalent to the voltage difference between $V_{IN}(+)$ and $V_{IN}(-)$. By reducing $V_{REF}(V_{REF}=V_{REF}(+)-V_{REF}(-))$ to less than 5V, the sensitivity of the converter can be increased (i.e., if $V_{REF}=2V$ then 1 LSB=7.8 mV). The input/reference arrangement also facilitates ratiometric operation and in many cases the chip power supply can be used for transducer power as well as the V_{REF} source.

This reference flexibility lets the input span not only be varied but also offset from zero. The voltage at $V_{REF}(-)$ sets the input level which produces a digital output of all zeroes. Though V_{IN} is not itself differential, the reference design affords nearly differential-input capability for most measurement applications. *Figure 9* shows some of the configurations that are possible.

2.2 INPUT CURRENT

Due to the unique conversion techniques employed by the ADC0820, the analog input behaves somewhat differently than in conventional devices. The A/D's sampled-data comparators take varying amounts of input current depending on which cycle the conversion is in.

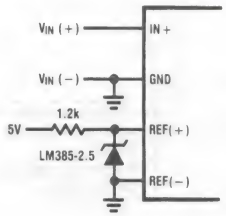
The equivalent input circuit of the ADC0820 is shown in *Figure 10a*. When a conversion starts ($\overline{WR}$ low, WR-RD mode), all input switches close, connecting V_{IN} to thirty-one 1 pF capacitors. Although the two 4-bit flash circuits are not both in their compare cycle at the same time, V_{IN} still sees all input capacitors at once. This is because the MS flash converter is connected to the input during its compare interval and the LS flash is connected to the input during its zeroing phase (Section 1.3). In other words, the LS ADC uses V_{IN} as its zero-phase input.

The input capacitors must charge to the input voltage through the on resistance of the analog switches (about 5 kΩ to 10 kΩ). In addition, about 12 pF of input stray capacitance must also be charged. For large source resistances, the analog input can be modeled as an RC network as shown in *Figure 10b*. As R_S increases, it will take longer for the input capacitance to charge.

In RD mode, the input switches are closed for approximately 800 ns at the start of the conversion. In WR-RD mode, the time that the switches are closed to allow this charging is the time that $\overline{WR}$ is low. Since other factors force this time to be at least 600 ns, input time constants of 100 ns can be accommodated without special consideration. Typical total input capacitance values of 45 pF allow R_S to be 1.5 kΩ without lengthening $\overline{WR}$ to give V_{IN} more time to settle.

2.0 Analog Considerations (Continued)

External Reference 2.5V Full-Scale



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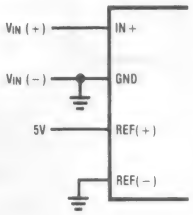
Power Supply as Reference



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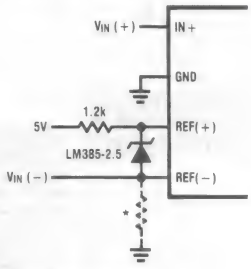
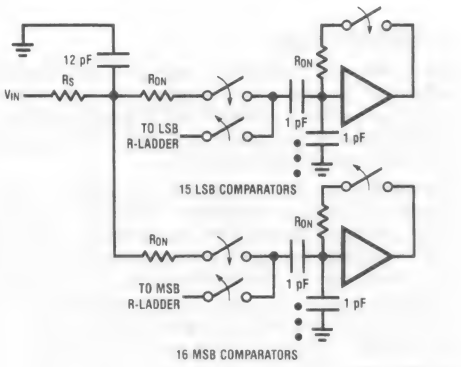
Input Not Referred to GND



* Current path must still exist from $V_{IN}(-)$ to ground

TL/H/5501–23

FIGURE 9. Analog Input Options



TL/H/5501–24

FIGURE 10a

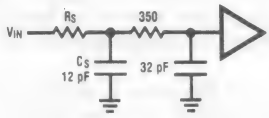


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FIGURE 10b

2.3 INPUT FILTERING

It should be made clear that transients in the analog input signal, caused by charging current flowing into V_{IN}, will not degrade the A/D's performance in most cases. In effect the ADC0820 does not "look" at the input when these transients occur. The comparators' outputs are not latched while $\overline{WR}$ is low, so at least 600 ns will be provided to charge the ADC's input capacitance. It is therefore not necessary to filter out these transients by putting an external cap on the V_{IN} terminal.

2.4 INHERENT SAMPLE-HOLD

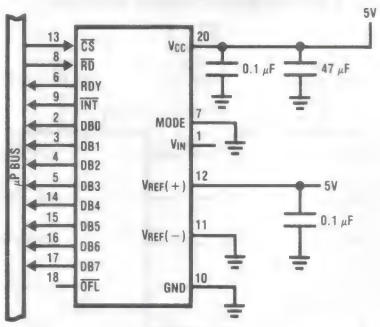
Another benefit of the ADC0820's input mechanism is its ability to measure a variety of high speed signals without the help of an external sample-and-hold. In a conventional SAR type converter, regardless of its speed, the input must remain at least ½ LSB stable throughout the conversion process if full accuracy is to be maintained. Consequently, for many high speed signals, this signal must be externally sampled, and held stationary during the conversion.

Sampled-data comparators, by nature of their input switching, already accomplish this function to a large degree (Section 1.2). Although the conversion time for the ADC0820 is 1.5 μs, the time through which V_{IN} must be 1/2 LSB stable is much smaller. Since the MS flash ADC uses V_{IN} as its "compare" input and the LS ADC uses V_{IN} as its "zero" input, the ADC0820 only "samples" V_{IN} when $\overline{WR}$ is low (Sections 1.3 and 2.2). Even though the two flashes are not done simultaneously, the analog signal is measured at one instant. The value of V_{IN} approximately 100 ns after the rising edge of $\overline{WR}$ (100 ns due to internal logic prop delay) will be the measured value.

Input signals with slew rates typically below 100 mV/μs can be converted without error. However, because of the input time constants, and charge injection through the opened comparator input switches, faster signals may cause errors. Still, the ADC0820's loss in accuracy for a given increase in signal slope is far less than what would be witnessed in a conventional successive approximation device. An SAR type converter with a conversion time as fast as 1 μs would still not be able to measure a 5V 1 kHz sine wave without the aid of an external sample-and-hold. The ADC0820, with no such help, can typically measure 5V, 7 kHz waveforms.

3.0 Typical Applications

8-Bit Resolution Configuration



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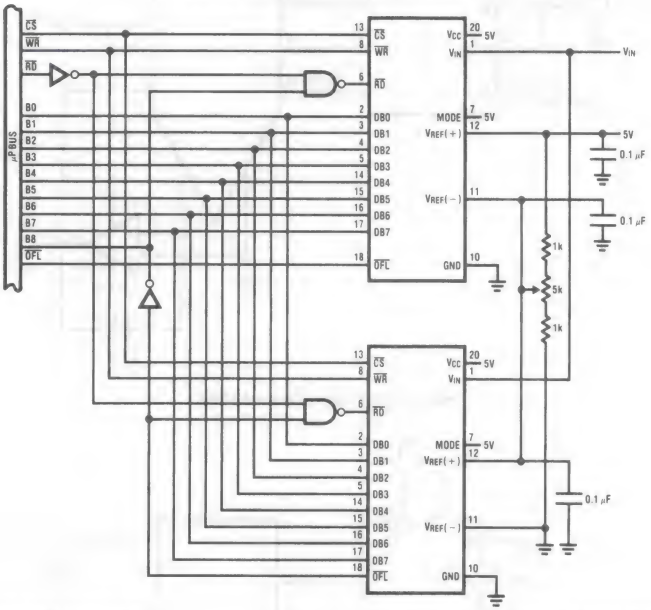
9-Bit Resolution Configuration



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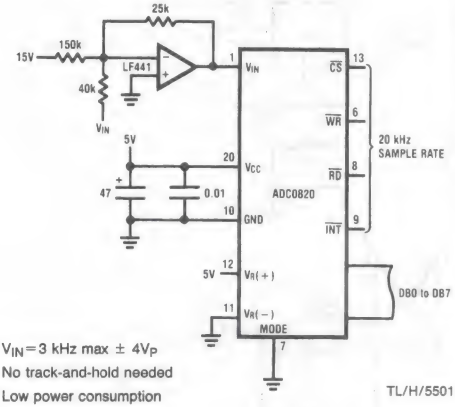
Telecom A/D Converter



- V_{IN} = 3 kHz max ± 4V_P
- No track-and-hold needed
- Low power consumption

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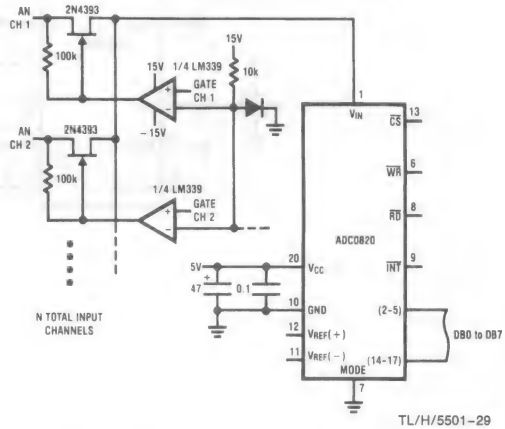
Multiple Input Channels



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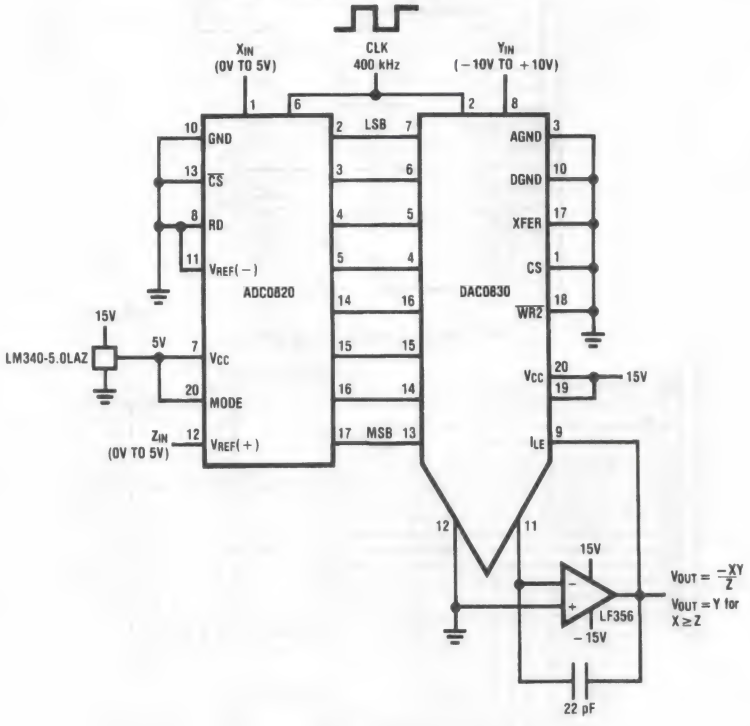
3.0 Typical Applications (Continued)

8-Bit 2-Quadrant Analog Multiplier



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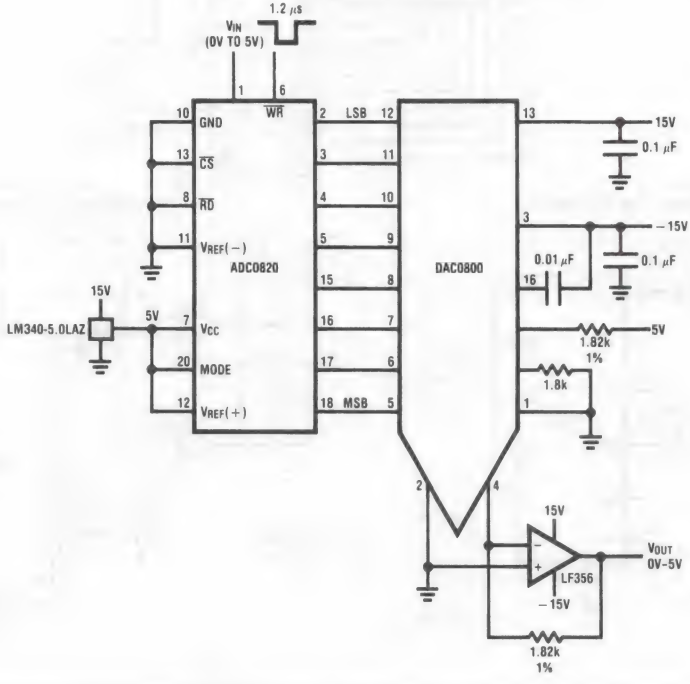
Fast Infinite Sample-and-Hold



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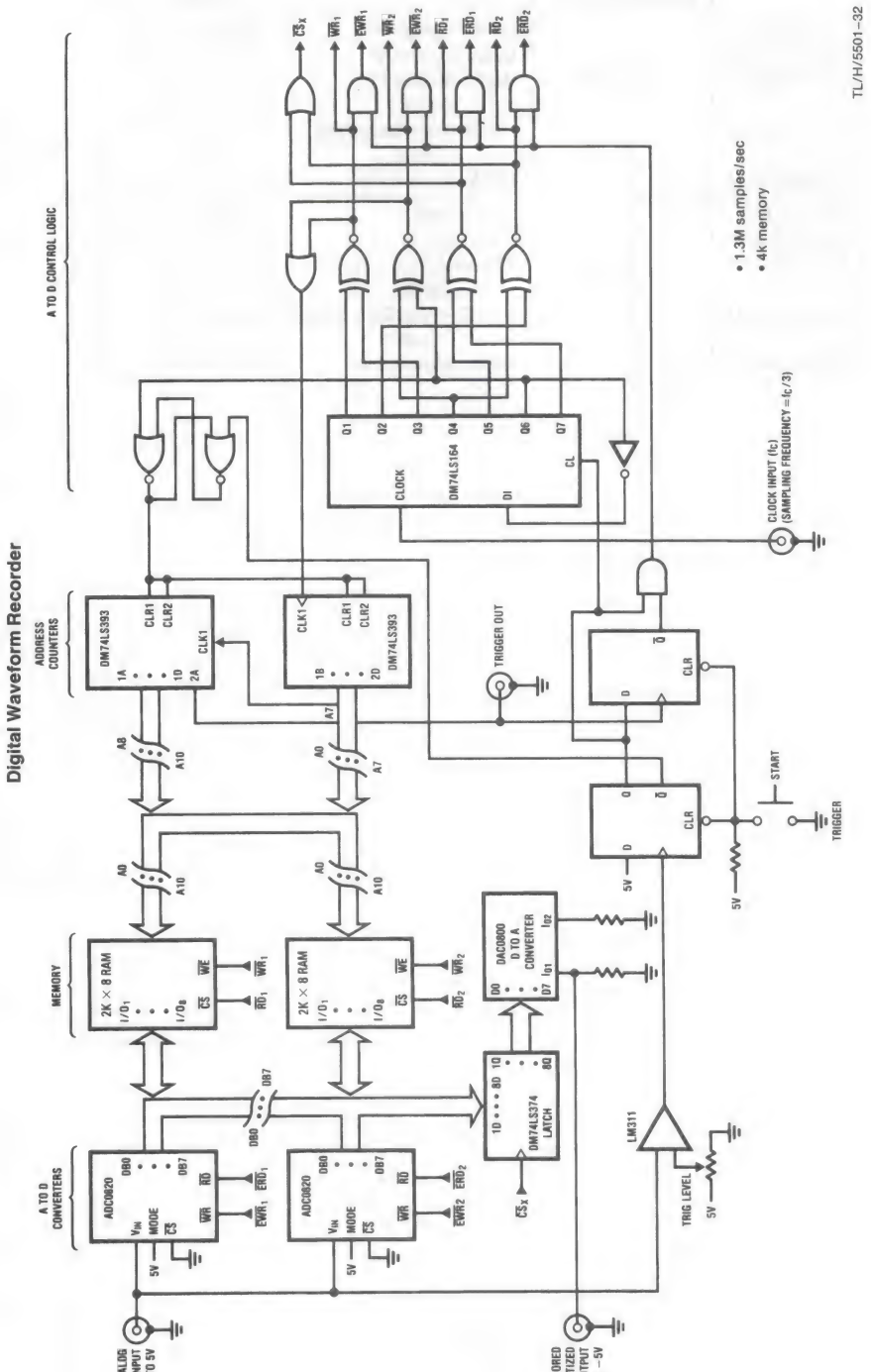
3.0 Typical Applications (Continued)



Ordering Information

Part Number	Total Unadjusted Error	Package	Temperature Range
ADC0820BD	±½ LSB	D20A—Cavity DIP	−55°C to +125°C
ADC0820BCD		D20A—Cavity DIP	−40°C to +80°C
ADC0820BCV		V20A—Molded Chip Carrier	0°C to +70°C
ADC0820BCM		M20B—Wide Body Small Outline	0°C to +70°C
ADC0820BCN		N20A—Molded DIP	0°C to +70°C
ADC0820CJ	±1 LSB	J20A—Cerdip	−55°C to +125°C
ADC0820CCJ		J20A—Cerdip	−40°C to +85°C
ADC0820CCV		V20A—Molded Chip Carrier	0°C to +70°C
ADC0820CCM		MJ20B—Wide Body Small Outline	0°C to +70°C
ADC0820CCN		N20A—Molded DIP	0°C to +70°C

National Semiconductor

ADC0829 μP Compatible 8-Bit A/D with 11-Channel MUX/Digital Input

General Description

The ADC0829 is an 8-bit successive approximation A/D converter with an 11-channel multiplexer of which six can be used as digital inputs, as well as, analog inputs.

This A/D is designed to operate from the μP data bus using a single 5V supply.

Channel selection, conversion control, software configuration and bus interface logic are all contained on this monolithic CMOS device.

This device contains three 16-bit registers which are accessed via double byte instructions. The control register is a write only register which controls the start of a new conversion, selects the channel to be converted, configures the 8-bit I/O port as input or output, and provides information for the 8-bit output register.

The conversion results register is a read only register which contains the current status and most recent conversion results. The discrete input register is also a read only register which contains the four address bits of the selected channel, and the six discrete inputs which are connected to the analog multiplexer.

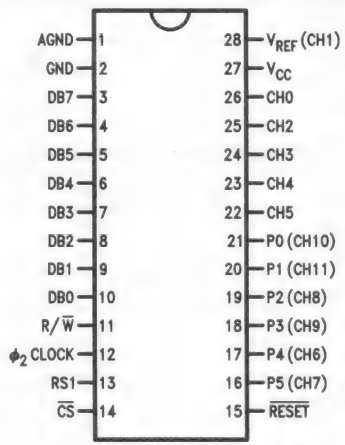
Features

- Easy interface to all microprocessors or operates "stand alone"
- Operates ratiometrically or with analog span adjusted voltage reference
- 11-Channel multiplexer with latched control logic of which six can be used as digital inputs
- 0 to 5V analog input range with single 5V supply
- TTL/MOS input/output compatible
- No zero or full scale adjusts required
- Standard 28-pin DIP
- Temperature range −40°C to +85°C
- ADC0829 equivalent to MM74C934

Key Specification

■ Resolution	8 Bits
■ Total Unadjusted Error	±½ LSB and ±1 LSB
■ Conversion Time	256 μs
■ Single Supply	$5V_{DC}$
■ Low Power	50 mW

Connection and Block Diagrams



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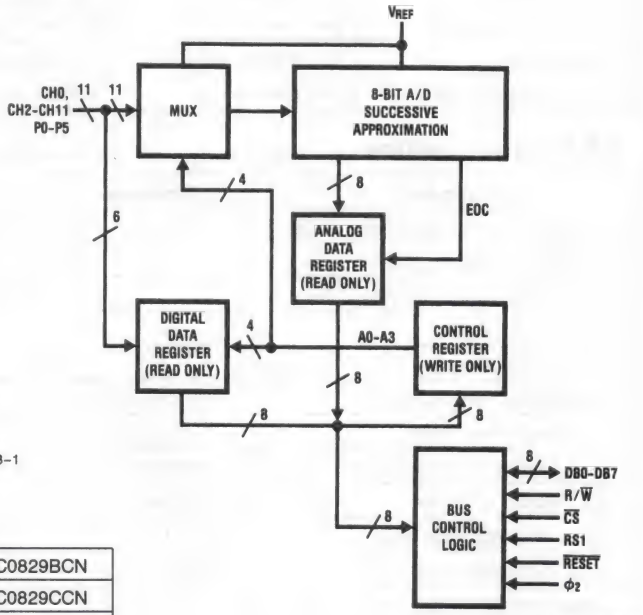
Top View



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Ordering Information

Error	± 1/2 Bit Unadjusted	ADC0829BCN
	± 1 Bit Unadjusted	ADC0829CCN
Package Outline		N28B

Absolute Maximum Ratings (Notes 1 and 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage, V_{CC} (Note 3)	6.5V
Voltage	
Logic Inputs	−0.3V to V_{CC} + 0.3V
Analog Inputs	−0.3V to V_{CC} + 0.3V
Storage Temperature	−65°C to + 150°C
Package Dissipation at T_A = 25°C (Board Mount)	875 mW
Lead Temp. (Soldering, 10 seconds)	260°C
ESD Susceptability (Note 8)	2000V
Input Current Per Pin	±5 mA
Package	+20 mA

Operating Conditions (Notes 1 and 2)

Supply Voltage, V_{CC}	4.75 V_{DC} to 5.5 V_{DC}
Temperature Range	−40°C to + 85°C

Converter and Multiplexer Electrical Characteristics

V_{CC} = 5V_{DC} = $V_{REF}(+)$, $V_{REF}(-)$ = GND, SCLK ϕ_2 = 1.048 MHz, −40°C ≤ T_A + 85°C unless otherwise noted.

Parameter	Conditions		Min	Typ (Notes)	Max	Units
Total Unadjusted Error; (Note 3)						
ADC0829BCN	V_{REF} Forced to 5.000 V_{DC}				±½	LSB
ADC0829CCN	V_{REF} Forced to 5.000 V_{DC}				±1	LSB
Reference Input Resistance			1.0	4.5		kΩ
Analog Input Voltage Range	(Note 4) V(+) or V(−)		GND-0.10		V_{CC}+0.10	V
$V_{REF}(+)$ Voltage, Top of Ladder	Measured at REF(+)			V_{CC}	V_{CC}+0.01	V
$\frac{V_{REF}(+) + V_{REF}(-)}{2}$ Voltage, Center of Ladder			$V_{CC/2}$−0.1	$V_{CC/2}$	$V_{CC/2}$+0.01	V
$V_{REF}(-)$ Voltage, Bottom of Ladder	Measured at REF(−)		−0.1	0		V
I_{OFF}, Off Channel	ON Channel = 5V	ADC0829BCN			±400	nA
Leakage Current (Note 6)	OFF Channel = 0V	ADC0829CCN			±1	μA
I_{ON}, On Channel	ON Channel = 0V	ADC0829BCN			±400	nA
Leakage Current (Note 6)	OFF Channel = 5V	ADC0829CCN			±1	μA

AC Characteristics

V_{CC} = $V_{REF}(+)$ = 5V, t_r = t_f = 20 ns and T_A = 25°C (Note 7) unless otherwise noted.

Parameter	Conditions	Min	Typ	Max	Units
$t_{CYC}(\phi_2)$, ϕ_2 Clock Cycle Time ($1/f_{\phi 2}$)		0.943		10.0	μs
$PW_H(\phi_2)$, ϕ_2 Clock Pulse Width, High		440			ns
$PW_L(\phi_2)$, ϕ_2 Clock Pulse Width, Low		410			ns
$t_r(\phi_2)$, ϕ_2 Rise Time				25	ns
$t_f(\phi_2)$, ϕ_2 Fall Time				30	ns
t_{AS}, Address Set Up Time	RS1, R/$\overline{W}$, $\overline{CS}$	145			ns
t_{DDR}, Data Delay (Read)	DB0-DB7			335	ns
t_{DSW}, Data Delay Setup (Write)	DB0-DB7	185			ns
t_{AH}, Address Hold Time	RS1, R/$\overline{W}$, $\overline{CE}$	20			ns
t_{DHW}, Input Data Hold Time	DB0-DB7	20			ns
t_{DHR}, Output Data Hold Time	DB0-DB7	10			ns
Analog Channel Settling Time		32			Clocks
t_c, Conversion Time		256			Clocks

Digital and DC Characteristics V_{CC} = 4.5V to 5.5V and $-40°C \leq T_A \leq 85°C$ unless otherwise noted.

Parameter	Conditions	Min	Typ	Max	Units
Bus Control Inputs (R/$\overline{W}$, ENABLE $\overline{RESET}$, RS1, $\overline{CS}$) and Peripheral Inputs (P0-P5)					
$V_{IN}(1)$, Logical "1" Input Voltage		2.0			V
$V_{IN}(0)$, Logical "0" Input Voltage				0.8	V
I_{IN}, Input Leakage Current				±1	μA
ϕ_2 CLOCK INPUT					
$V_{IN}(1)$, Logical "1" Input Voltage		$V_{CC}-0.8$			V
$V_{IN}(0)$, Logical "0" Input Voltage				0.4	V
Data Bus (DB0-DB7)					
$V_{IN}(1)$, Logical "1" Input Voltage		2.0			V
$V_{IN}(0)$, Logical "0" Input Voltage				0.8	V
I_{OUT}, TRI-STATE® Output Current	$V_{OUT}=0V$			−10	μA
	$V_{OUT}=5V$			10	μA
$V_{OUT}(1)$, Logical "1" Output Voltage	$I_{OUT}=-1.6$ mA	2.4			V
$V_{OUT}(0)$, Logical "0" Output Voltage	$I_{OUT}=1.6$ mA			0.4	V
Power Supply Requirements					
I_{CC}, Supply Current				10	mA

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to ground.

Note 3: Total unadjusted error includes offset, full-scale, linearity, and multiplexer error.

Note 4: For $V_{IN}(-) \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input, which will forward-conduct for analog input voltages one diode drop below ground or one diode drop greater than V_{CC} supply. Be careful during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct, especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 100 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 100 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.90 V_{DC} over temperature variations, initial tolerance and loading.

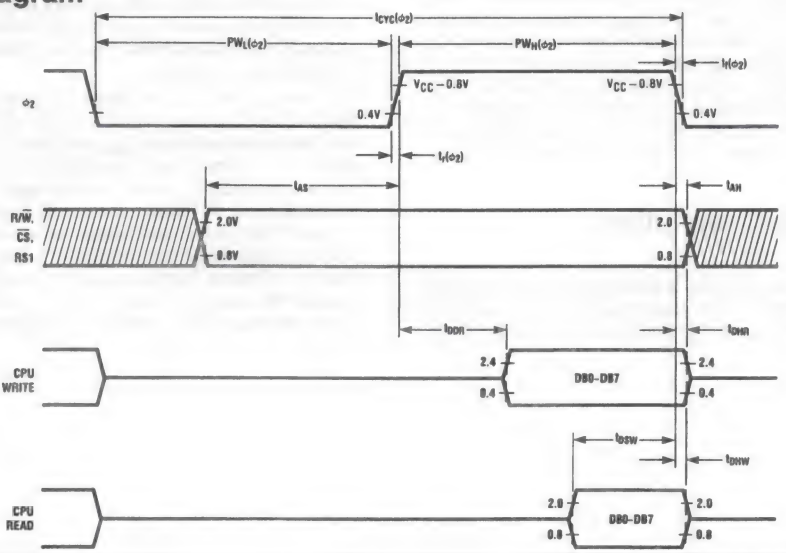
Note 5: Typicals are at 25°C and represent most likely parametric norm.

Note 6: Off channel leakage current is measured after the channel selection.

Note 7: The temperature coefficient is 0.3%/°C.

Note 8: Human Body Model, 100 pF discharged through a 1.5 kΩ resistor.

Timing Diagram



TL/H/5508–3

Pin Descriptions

ANALOG AND DIGITAL INPUTS

CH0, CH2-CH5—These are dedicated analog inputs. They are fed directly to the internal 12 to 1 multiplexer which feeds the A/D converter.

P0-P5/CH6-CH11—These 6 pins are dual purpose and may be used as either TTL compatible digital inputs, or analog inputs. When used as digital inputs they may be read via the discrete input register. When they are used as analog inputs they function like CH-0, CH2-5.

MICROPROCESSOR INTERFACE SIGNALS

DB0-DB7—The bi-directional data lines for the data bus connect to the μP's main data bus to enable data transfer to and from the μP. DB0-DB7 remain in a high impedance state unless the ADC0829 is read.

ϕ_2 Clock—This signal is used for two purposes. First it synchronizes data transfer in and out of the ADC. Second, it is the master clock for the A/D converter logic and all other timing signals are derived from it.

R/$\overline{W}$—The read/write pin controls the direction of data transfer on D0-D7.

$\overline{RESET}$—A low on this pin forces the ADC0829 into a known state. The start bit is cleared, Channel CH0 is selected and the internal byte counter is reset to the MS Byte. The A/D data register is not reset. Reset must be held low for at least 3 clocks.

$\overline{CS}$—Chip Select must be low in order for data transfer between the ADC0829 and the μP to occur.

RS1—The Register Select pin is used to address the internal registers.

POWER SUPPLY PINS

V_{CC}—This is the positive 5V supply pin. It powers the digital load and the sample data comparator. Care should be exercised to ensure that supply noise on this pin is adequately filtered, by using a bypass capacitor from V_{CC} to D_{GND}.

D_{GND}—Digital ground should be connected to the systems digital ground.

V_{REF} and A_{GND}—The positive reference pin attaches to the top of the 256R resistor ladder and sets the full scale conversion voltage value. The A_{GND} connects to the bottom of the ladder. The conversion result is ratiometric to V_{REF} - A_{GND} and hence both V_{REF} and A_{GND} should be noise free. Ideally the V_{REF} and A_{GND} should be single point connected to the analog transducer's supply. The V_{REF} and A_{GND} voltages typically are 5V and Ground but they may be varied so long as $(V_{REF}-A_{GND})/2 = V_{CC}/2 \pm 0.1V$.

Functional Description

1.0 CONTROL LOGIC

The Control Logic interprets the microprocessor control signals and decodes these signals to perform the actual functions of selecting, reading, writing, enabling the outputs, etc.

2.0 STATE DESCRIPTIONS

There are three internal states within the A/D converter: the NO OP state; the sample state; and the converting state.

The NO OP state is a stable state since the external stimulus (e.g. start conversion signal) is needed for a state transition.

The first transient state is sampling the input. The first 32 clocks of the conversion are used for acquiring the channel; this settling time allows any transients to decay before conversion begins. The second transient state is the actual conversion. The conversion is completed in 256 clocks and the conversion results register is updated. The converter then returns to the stable NO OP state awaiting further instructions.

The device has no comparator bias current and draws minimal power during the NO OP state.

3.0 INITIALIZATION

The device is initialized by an active low on $\overline{RESET}$. All outputs are initialized to the inactive state and the converter placed in its NO OP state. The data register is not affected by $\overline{RESET}$. System TRI-STATE outputs are initialized to the high impedance state.

4.0 CONVERSION CONTROL

The program normally initiates a conversion cycle with a double write command. (See control word format.) The control word selects a channel, configures the peripheral I/O, and provides peripheral data information. The conversion is initiated by setting the SC bit in the control word high.

The converter then resets the start conversion bit and begins the conversion cycle.

When the conversion is complete and the new conversion results transferred to the data register, the status bit is set. The status bit is not reset when the conversion status is read. A full double byte write into the control word will reset the status bit, or a low level at master $\overline{RESET}$.

If a new conversion command occurs during a conversion, the conversion is aborted and a new channel acquisition phase will immediately begin.

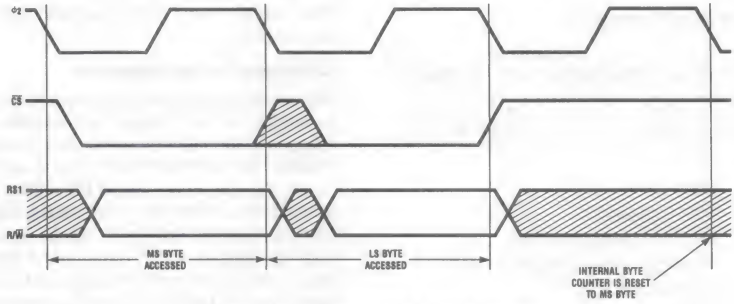
5.0 CONTROL STRUCTURE

The control logic continually monitors the control bus waiting for $\overline{CS}$ to go low and ϕ_2 to go high. When this condition occurs, the internal decoder, which has already selected the proper function, activitates.

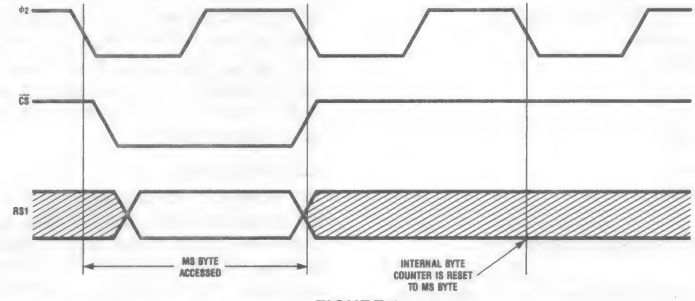
The byte counter will always select the most significant (MS) half first, and the least significant (LS) half second. Single byte instructions will always access the MSB portion of any word. After a single byte instruction the byte counter will return to the MSB portion of a word when $\overline{CS}$ is high for a full clock cycle. A 16-bit read or write is accomplished by using a 16-bit load or store instruction which transfers each byte on consecutive clock cycles. This timing is shown in *Figure 1*. A single byte instruction is especially useful for reading the status bit during a polled interrupt. *Figure 2* shows the basic A/D conversion timing sequence and flow.

Functional Description (Continued)

Timing for a Typical μP 16 Byte Access

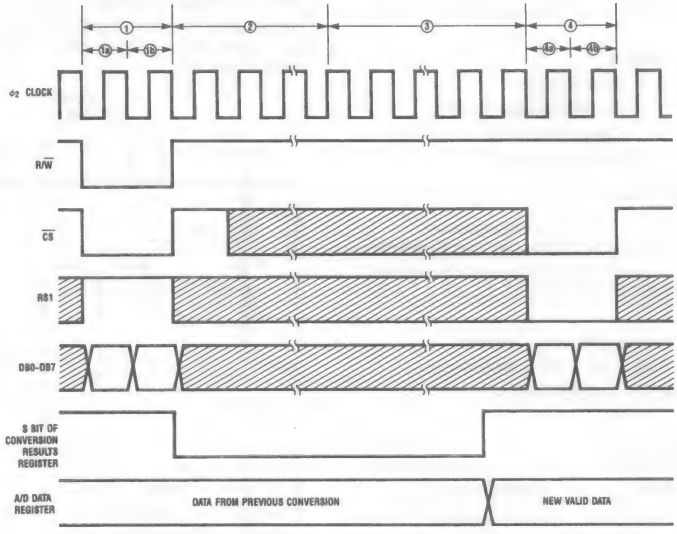


Timing for a Typical μP 8 Byte Access



TL/H/5508–4

FIGURE 1



(1) START CONVERSION
- (a) SET SC BIT TO A 1
- (b) LOAD ADDRESS

(2) ANALOG INPUT SETTLING TIME ALLOWS INTERNAL MULTIPLEXER TO SELECT A CHANNEL AND STABILIZE (~32 CLOCKS).

(3) A/D CONVERSION TIME (~256 CLOCKS)

(4) READ END OF CONVERSION DATA
- (a) EOC BIT READ IF A 1 CONVERSION COMPLETE.
- (b) A/D DATA REGISTER READ. IF EOC = 1, THEN NEW VALID DATA.

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FIGURE 2. A/D Conversion Timing Sequence

Functional Description (Continued)

6.0 WORD FORMAT

6.1 Control Register Word Format

← MSB Word →								← LSB WORD →							
DB_7	DB_6	DB_5	DB_4	DB_3	DB_2	DB_1	DB_0	DB_7	DB_6	DB_5	DB_4	DB_3	DB_2	DB_1	DB_0
X	X	X	X	X	X	X	(LSB) SC	X	X	X	X	A3 CH3	A2 CH2	A1 CH1	A0 CH0

X: Don't Care

SC: Start Conversion
- 1 = Start new conversion
- 0 = Do not start new conversion

CH3-CH0: Channel Address

Hex Value	Definition
0	Select CH0
1	Select $V_{ref}(+)$
2-5	Select Channels CH2-CH5
6-9	Undefined
A	CH10
B	CH11
C	CH8
D	CH9
E	CH6
F	CH7

6.2 Conversion Results Register Word Format

← MSB Word →								← LSB WORD →							
DB_7	DB_6	DB_5	DB_4	DB_3	DB_2	DB_1	DB_0	DB_7	DB_6	DB_5	DB_4	DB_3	DB_2	DB_1	DB_0
S	0	0	0	0	0	0	0	C_7	C_6	C_5	C_4	C_3	C_2	C_1	C_0

S: Status
- 1 = Data is valid (conversion complete)
- 0 = Data is not valid

C_7-C_0: 8 bit converted result

6.3 Discrete Input Word Format

← MSB Word →								← LSB WORD →							
DB_7	DB_6	DB_5	DB_4	DB_3	DB_2	DB_1	DB_0	DB_7	DB_6	DB_5	DB_4	DB_3	DB_2	DB_1	DB_0
P5	P4	P3	P2	P1	P0	CH3	CH2	CH1	CH0	0	0	0	0	0	0

CH3-CH0: Status of channel address

P5-P0: Status of P5-P0 interpreted as discrete digital inputs

ADU ADDRESS SELECTION

CSO*	R/W	RSI	Description
1	X	X	Do not respond
0	0	0	Write NO OP
0	0	1	Write Control Word
0	1	0	Read Conversion Results
0	1	1	Read Discrete Inputs

Note: All words are transferred as two 8-bit bytes, MSB transferred first LSB transferred second.

7.0 ANALOG TO DIGITAL CONVERTER

The ADC0829 A/D Converter is composed of three major sections: the successive approximation register (SAR); the 256R ladder and analog decoder; and the sample-data comparator.

7.1 Successive Approximation

The analog signal at the A/D input is compared eight times to various ladder voltages to determine which of the 256 voltages in the ladder most closely approximates the input voltage. This stochastic technique is accomplished by converging on the proper tap in the ladder by simple iterative convergence. There are nine posting registers in the SAR which contain the position of the bit being tested and eight latching registers which remember if the comparison was high or low. Starting with the MSB and continuing downward each bit is set high by the posting register. The analog tree decoder selects the corresponding tap in the ladder and the A/D input is compared to that voltage. If the comparison is positive the latch remains set, so higher voltages in the ladder are checked next. If the comparison is negative the bit is reset so lower ladder voltages are sought.

After all eight comparisons are made, the contents of the latching register are transferred to a data register, thus the A/D can perform a new conversion while the previous results remain available.

7.2 256R Ladder

The ladder is a very accurate voltage divider which divides the reference voltage into 256 equal steps. Special consideration was given to the ladder terminations at each end, and also the center, to ensure consistent and accurate voltage steps. The use of a 256R ladder guarantees monotonicity since only a single voltage gradient across the ladder exists. Shorted or unequal resistors in the ladder may cause non-uniform steps but cannot cause a nonmonotonic response so often fatal in closed loop system applications. (See *Figure 3.*)

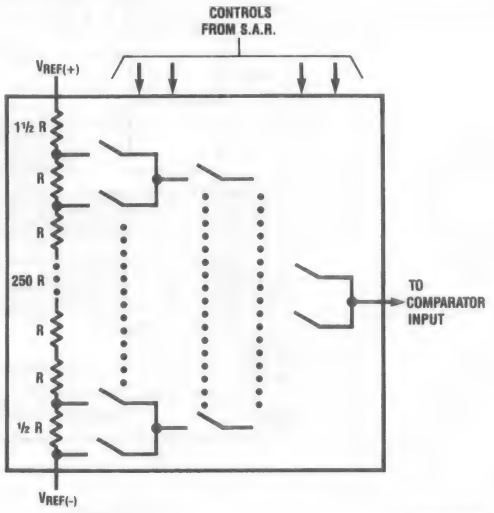

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FIGURE 3. Resistor Ladder and Switch Tree

Functional Description (Continued)

Actually of the 256 resistors in the ladder, 254 have the same value while the end point resistors are equal to 1-1/2R and 1/2R. This ensures the system output characteristic is symmetrical with the zero and full scale points of its input to output, or transfer curve.

The tree decoder routes the 256 voltages from the ladder to a single point at the comparator input. This allows comparisons between the A/D input and any voltage the SAR directs the decoder to route to the comparator.

Since the ladder is dependent upon only the matching of resistors, the voltages it generates are very stable with temperature and have excellent repeatability and long term drift.

8.0 MULTIPLEXER

8.1 Analog Inputs

The analog multiplexer selects one of 11 channels and directs them to the input of the A/D converter. The multiplexer was designed to minimize the effects of leakage currents and multiplexer output capacitance.

Special input protection is used to prevent damage from static voltages or voltages exceeding the specified range from $-0.3V$ to $V_{CC}+0.3V$. However, normal precautions are recommended to avoid such situations whenever possible.

8.2 Digital Inputs

Six of the analog inputs can also be used as digital inputs to sense TTL voltage levels. Care must be taken when these inputs are interpreted since TTL levels may not always be present.

8.3 A/D Comparator

Probably the most important section of the A/D converter is the comparator since the comparator's offset voltage and stability determine the converter's ultimate accuracy. The low voltage offset of the chopper-stabilized comparator of this converter optimizes performance by minimizing temperature dependent input offset errors as well as drift.

The dc signal appearing at the amplifier input is converted to an ac signal, amplified by an ac amplifier and restored to a dc signal. The drift of the comparator is minimized since the drift signal is a dc component blocked by the ac amplifier.

The comparator has very high input impedance to dc voltages since it looks like a capacitor. Because the comparator is chopping the dc voltages at the input, the difference between the A/D input voltage and ladder voltage appears on the comparator's input capacitor. The input voltage difference, chopping frequency, and comparator input capacitor causes a CVF current. The CVF current is a small bias current which will not produce any error when the A/D input is connected to a low impedance voltage source. If the voltage source has an output impedance of less than 10k, the error is still insignificant since the bias current exponentially decays.

Adding a capacitor to the input of the comparator integrates the exponential charging current converting it into dc bias current. (See Figure 1.) Two main considerations on the integration capacitor are charge sharing with a filter capacitor and settling time.

9.0 BUS INTERFACE

The ADC0829 communicates to the microprocessor through an 8-bit I/O port. The I/O port is composed of a TTL to CMOS buffer and a TRI-STATE® output driver.

The TTL to CMOS Buffer translates the TTL voltage levels into CMOS levels very rapidly and is quite stable with supply and temperature. The buffer has a small amount of hysteresis (about 100 mV) to improve both noise immunity and internal rise and fall times.

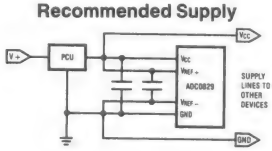
The TRI-STATE bus driver is a bipolar and N-channel pair that easily drive the bus capacitance. Since the bus drivers collectively can sink or source a quarter of an amp total, a non-overlap circuit is used which guarantees that only one of the two drive transistors is on at a time.

Since this output drives the bus capacitance, even the non-overlapping circuit cannot prevent noise on V_{CC}. The amount of noise depends on the V_{CC} current used to charge the bus capacitance.

The external filter capacitor on V_{CC} provides some of the transient current while the bus is being driven. A capacitor with good ac characteristics and low series resistance is a good choice to prevent V_{CC} transients from affecting accuracy.

Application Information

Recommended Supply



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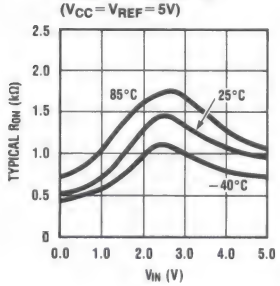
Multiplexer R_{ON} vs V_{IN} ($V_{CC}=V_{REF}=5V$)



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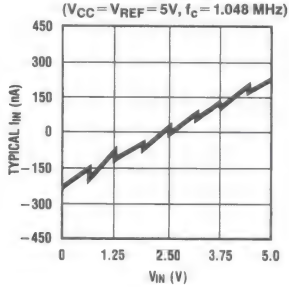
Comparator I_{IN} vs V_{IN} ($V_{CC}=V_{REF}=5V$, $f_C=1.048$ MHz)



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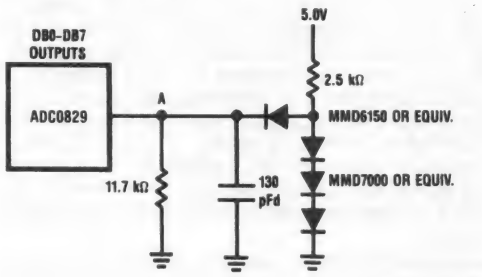
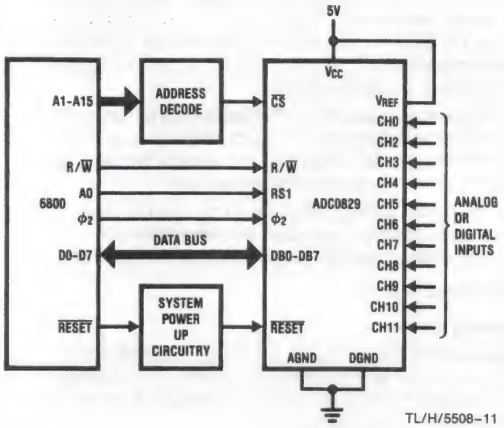
Data Bus Test Circuit



TL/H/5508–10

Typical Application



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National Semiconductor

ADC0831/ADC0832/ADC0834 and ADC0838 8-Bit Serial I/O A/D Converters with Multiplexer Options

General Description

The ADC0831 series are 8-bit successive approximation A/D converters with a serial I/O and configurable input multiplexers with up to 8 channels. The serial I/O is configured to comply with the NSC MICROWIRE™ serial data exchange standard for easy interface to the COPS™ family of processors, and can interface with standard shift registers or μPs.

The 2-, 4- or 8-channel multiplexers are software configured for single-ended or differential inputs as well as channel assignment.

The differential analog voltage input allows increasing the common-mode rejection and offsetting the analog zero input voltage value. In addition, the voltage reference input can be adjusted to allow encoding any smaller analog voltage span to the full 8 bits of resolution.

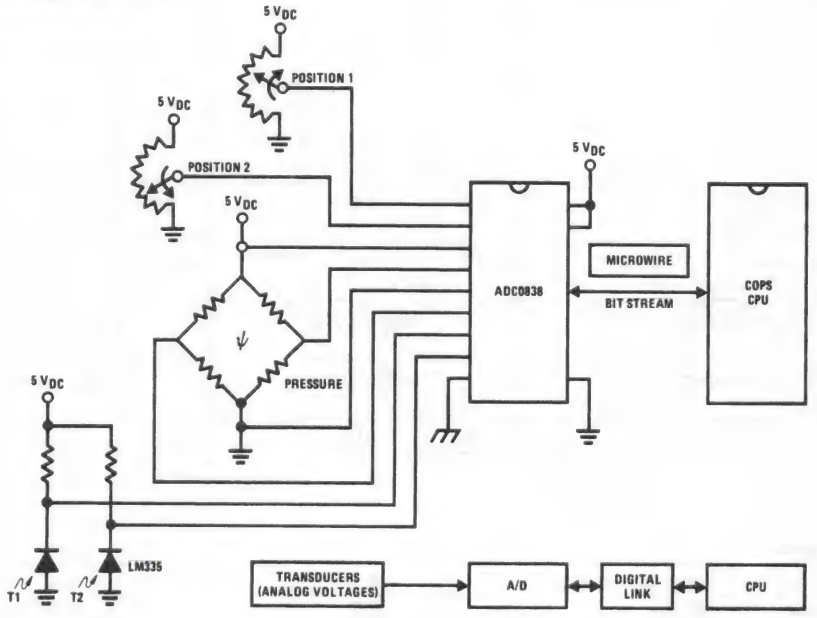
Features

- NSC MICROWIRE compatible—direct interface to COPS family processors
- Easy interface to all microprocessors, or operates "stand-alone"
- Operates ratiometrically or with 5 V_{DC} voltage reference
- No zero or full-scale adjust required
- 2-, 4- or 8-channel multiplexer options with address logic
- Shunt regulator allows operation with high voltage supplies
- 0V to 5V input range with single 5V power supply
- Remote operation with serial digital data link
- TTL/MOS input/output compatible
- 0.3" standard width, 8-, 14- or 20-pin DIP package
- 20 Pin Molded Chip Carrier Package (ADC0838 only)

Key Specifications

■ Resolution	8 Bits
■ Total Unadjusted Error	±½ LSB and ±1 LSB
■ Single Supply	5 V_{DC}
■ Low Power	15 mW
■ Conversion Time	32 μs

Typical Application



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Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Current into V+ (Note 3)	15 mA
Supply Voltage, V_{CC} (Note 3)	6.5V
Voltage	
Logic Inputs	−0.3V to V_{CC} + 0.3V
Analog Inputs	−0.3V to V_{CC} + 0.3V
Input Current per Pin (Note 4)	±5 mA
Package	±20 mA
Storage Temperature	−65°C to +150°C
Package Dissipation at T_A = 25°C (Board Mount)	0.8W
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C
ESD Susceptibility (Note 5)	2000V

Operating Ratings (Notes 1 & 2)

Supply Voltage, V_{CC}	4.5 V_{DC} to 6.3 V_{DC}
Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0831/2/4/8BJ ADC0831/2/4/8CJ	−55°C to +125°C
ADC0831/2/4/8BCJ ADC0831/2/4/8CCJ	−40°C to +85°C
ADC0831/2/4/8BCN ADC0838BCV ADC0831/2/4/8CCN ADC0838CCV	−0°C to +70°C

Converter and Multiplexer Electrical Characteristics

The following specifications apply for $V_{CC} = V+ = V_{REF} = 5V$, $V_{REF} \le V_{CC} + 0.1V$, $T_A = T_j = 25°C$, and $f_{CLK} = 250$ kHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX}.**

Parameter	Conditions	BJ, CJ, BCJ and CCJ Devices			BCV, CCV, BCN and CCN Devices			Units
		Typ (Note 12)	Tested Limit (Note 13)	Design Limit (Note 14)	Typ (Note 12)	Tested Limit (Note 13)	Design Limit (Note 14)	
CONVERTER AND MULTIPLEXER CHARACTERISTICS								
Total Unadjusted Error	V_{REF} = 5.00 V							
ADC0838BCV	(Note 6)					±½	±½	
ADC0831/2/4/8BCN							**±½**	LSB
ADC0831/2/4/8BJ			**±½**			±½		
ADC0831/2/4/8BCJ			**±½**					
ADC0838CCV						±1	±1	
ADC0831/2/4/8CCN						±1	**±1**	
ADC0831/2/4/8CJ			**±1**					
ADC0831/2/4/8CCJ			**±1**					
Minimum Reference Input Resistance (Note 7)		3.5	**1.3**		3.5	1.3	**1.3**	kΩ
Maximum Reference Input Resistance (Note 7)		3.5	**5.9**		3.5	5.4	**5.9**	kΩ
Maximum Common-Mode Input Range (Note 8)			**V_{CC} + 0.05**			V_{CC} + 0.05	**V_{CC} + 0.05**	V
Minimum Common-Mode Input Range (Note 8)			**GND − 0.05**			GND − 0.05	**GND − 0.05**	V
DC Common-Mode Error		±1/16	**±¼**		±1/16	±¼	**±¼**	LSB
Change in zero error from V_{CC} = 5V to internal zener operation (Note 3)	15 mA into V+ V_{CC} = N.C. V_{REF} = 5V		**1**			1	**1**	LSB
V_Z, internal diode breakdown (at V+) (Note 3) MIN	15 mA into V+		**6.3**			6.3	**6.3**	
MAX			**8.5**			8.5	**8.5**	V
Power Supply Sensitivity	V_{CC} = 5V ± 5%	±1/16	**±¼**	±¼	±1/16	±¼	**±¼**	LSB

Converter and Multiplexer Electrical Characteristics (Continued)

The following specifications apply for V_{CC} = V+ = 5V, T_A = T_j = 25°C, and f_{CLK} = 250 kHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX}.**

Parameter	Conditions	BJ, CJ, BCJ and CCJ Devices			BCV, CCV, BCN and CCN Devices			Units
		Typ (Note 12)	Tested Limit (Note 13)	Design Limit (Note 14)	Typ (Note 12)	Tested Limit (Note 13)	Design Limit (Note 14)	
CONVERTER AND MULTIPLEXER CHARACTERISTICS (Continued)								
I_{OFF}, Off Channel Leakage Current (Note 9)	On Channel = 5V, Off Channel = 0V		−0.2 **−1**			−0.2	**−1**	μA
	On Channel = 0V, Off Channel = 5V		+0.2 **+1**			+0.2	**+1**	μA
I_{ON}, On Channel Leakage Current (Note 9)	On Channel = 0V, Off Channel = 5V		−0.2 **−1**			−0.2	**−1**	μA
	On Channel = 5V, Off Channel = 0V		+0.2 **+1**			+0.2	**+1**	μA
DIGITAL AND DC CHARACTERISTICS								
$V_{IN(1)}$, Logical "1" Input Voltage (Min)	V_{CC} = 5.25V		**2.0**			2.0	**2.0**	V
$V_{IN(0)}$, Logical "0" Input Voltage (Max)	V_{CC} = 4.75V		**0.8**			0.8	**0.8**	V
$I_{IN(1)}$, Logical "1" Input Current (Max)	V_{IN} = 5.0V	0.005	**1**		0.005	1	**1**	μA
$I_{IN(0)}$, Logical "0" Input Current (Max)	V_{IN} = 0V	−0.005	**−1**		−0.005	−1	**−1**	μA
$V_{OUT(1)}$, Logical "1" Output Voltage (Min)	V_{CC} = 4.75V I_{OUT} = −360 μA I_{OUT} = −10 μA		 **2.4** **4.5**			 2.4 4.5	 **2.4** **4.5**	 V V
$V_{OUT(0)}$, Logical "0" Output Voltage (Max)	V_{CC} = 4.75V I_{OUT} = 1.6 mA		**0.4**			0.4	**0.4**	V
I_{OUT}, TRI-STATE Output Current (Max)	V_{OUT} = 0V V_{OUT} = 5V	−0.1 0.1	**−3** **3**		−0.1 0.1	−3 +3	**−3** **+3**	μA μA
I_{SOURCE}, Output Source Current (Min)	V_{OUT} = 0V	−14	**−6.5**		−14	−7.5	**−6.5**	mA
I_{SINK}, Output Sink Current (Min)	V_{OUT} = V_{CC}	16	**8.0**		16	9.0	**8.0**	mA
I_{CC}, Supply Current (Max) ADC0831, ADC0834, ADC0838		0.9	**2.5**		0.9	2.5	**2.5**	mA
ADC0832	Includes Ladder Current	2.3	**6.5**		2.3	6.5	**6.5**	mA

3

AC Characteristics

The following specifications apply for V_{CC} = 5V, $t_r = t_f$ = 20 ns and 25°C unless otherwise specified.

Parameter		Conditions	Typ (Note 12)	Tested Limit (Note 13)	Design Limit (Note 14)	Limit Units
f_{CLK}, Clock Frequency	Min			10		kHz
	Max				400	kHz
t_C, Conversion Time		Not including MUX Addressing Time		8		$1/f_{CLK}$
Clock Duty Cycle (Note 10)	Min				40	%
	Max				60	%
t_{SET-UP}, $\overline{CS}$ Falling Edge or Data Input Valid to CLK Rising Edge					250	ns
t_{HOLD}, Data Input Valid after CLK Rising Edge					90	ns
t_{pd1}, t_{pd0}—CLK Falling Edge to Output Data Valid (Note 11)		C_L = 100 pF				
		Data MSB First	650		1500	ns
		Data LSB First	250		600	ns
t_{1H}, t_{0H},—Rising Edge of CS to Data Output and SARS Hi–Z		C_L = 10 pF, R_L = 10k (see TRI-STATE® Test Circuits)	125		250	ns
		C_L = 100 pf, R_L = 2k		500		ns
C_{IN}, Capacitance of Logic Input			5			pF
C_{OUT}, Capacitance of Logic Outputs			5			pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to the ground plugs.

Note 3: Internal zener diodes (6.3 to 8.5V) are connected from V+ to GND and V_{CC} to GND. The zener at V+ can operate as a shunt regulator and is connected to V_{CC} via a conventional diode. Since the zener voltage equals the A/D's breakdown voltage, the diode insures that V_{CC} will be below breakdown when the device is powered from V+. Functionality is therefore guaranteed for V+ operation even though the resultant voltage at V_{CC} may exceed the specified Absolute Max of 6.5V. It is recommended that a resistor be used to limit the max current into V+. (See *Figure 3* in Functional Description Section 6.0)

Note 4: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 5 mA or less. The 20 mA package input current limits the number of pins that can exceed the power supply boundaries with a 5 mA current limit to four.

Note 5: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 6: Total unadjusted error includes offset, full-scale, linearity, and multiplexer errors.

Note 7: Cannot be tested for ADC0832.

Note 8: For $V_{IN}(-) \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input (see Block Diagram) which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater then the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct—especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

Note 9: Leakage current is measured with the clock not switching.

Note 10: A 40% to 60% clock duty cycle range insures proper operation at all clock frequencies. In the case that an available clock has a duty cycle outside of these limits, the minimum, time the clock is high or the minimum time the clock is low must be at least 1 μs. The maximum time the clock can be high is 60 μs. The clock can be stopped when low so long as the analog input voltage remains stable.

Note 11: Since data, MSB first, is the output of the comparator used in the successive approximation loop, an additional delay is built in (see Block Diagram) to allow for comparator response time.

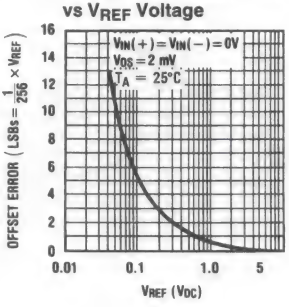
Note 12: Typicals are at 25°C and represent most likely parametric norm.

Note 13: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

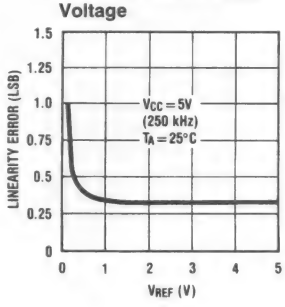
Note 14: Guaranteed but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Typical Performance Characteristics

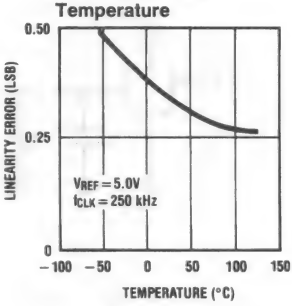
Unadjusted Offset Error vs V_{REF} Voltage



Linearity Error vs V_{REF} Voltage



Linearity Error vs Temperature



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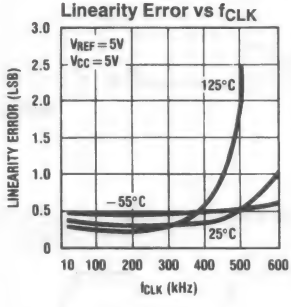
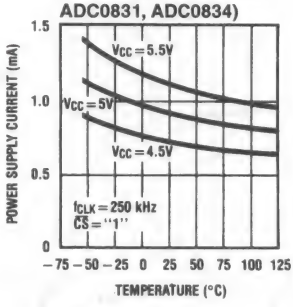
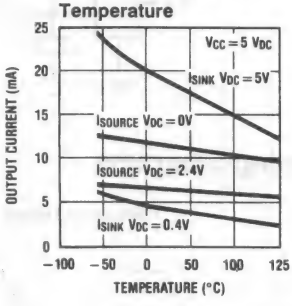
Linearity Error vs f_{CLK}



Power Supply Current vs Temperature (ADC0838, ADC0831, ADC0834)



Output Current vs Temperature



Note: For ADC0832 add I_{REF}.

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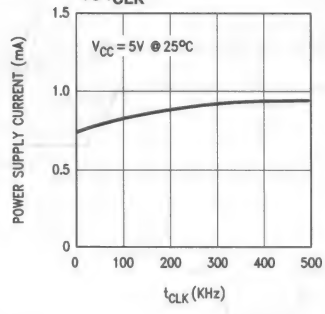
Power Supply Current vs f_{CLK}



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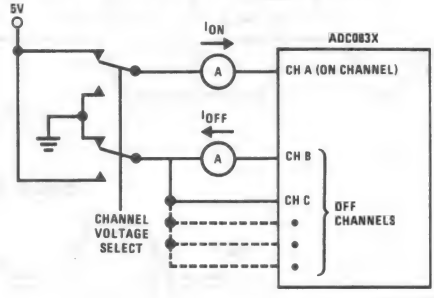
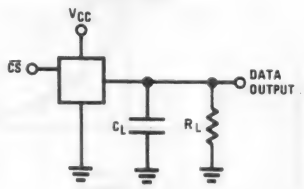
Leakage Current Test Circuit



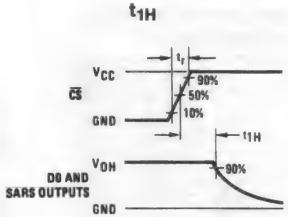
TL/H/5583–3

TRI-STATE Test Circuits and Waveforms

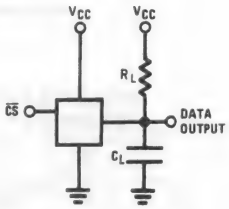
t_{1H}



t_{1H}



t_{0H}



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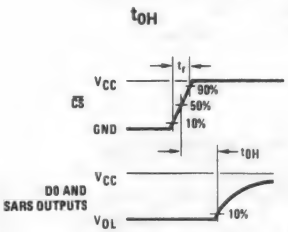
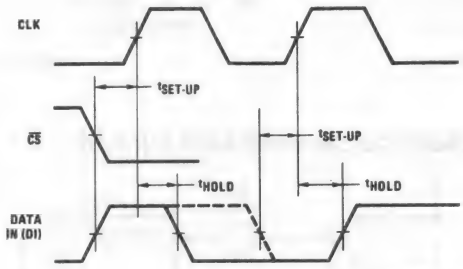
t_{0H}



TL/H/5583–23

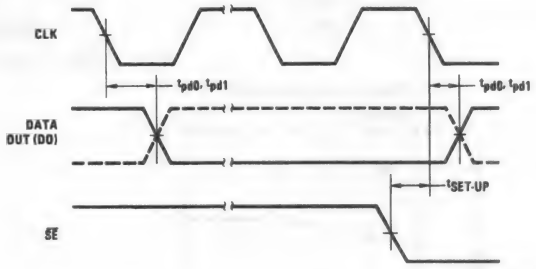
Timing Diagrams

Data Input Timing



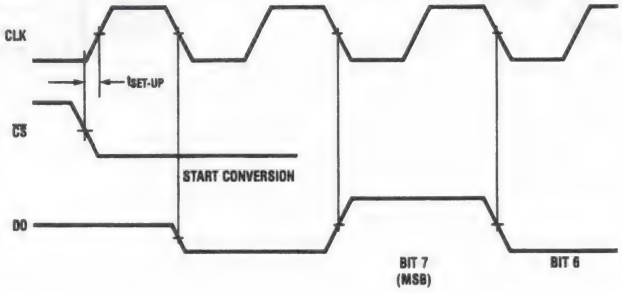
TL/H/5583–24

Data Output Timing



TL/H/5583–25

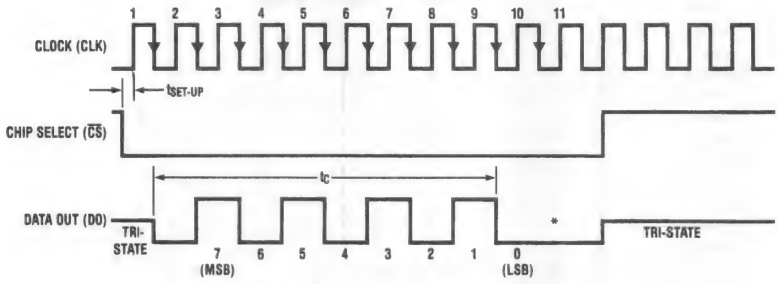
ADC0831 Start Conversion Timing



TL/H/5583–26

Timing Diagrams (Continued)

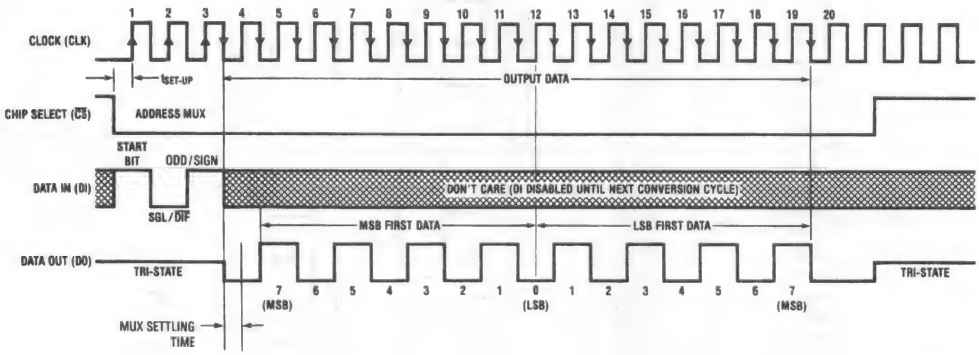
ADC0831 Timing



TL/H/5583–27

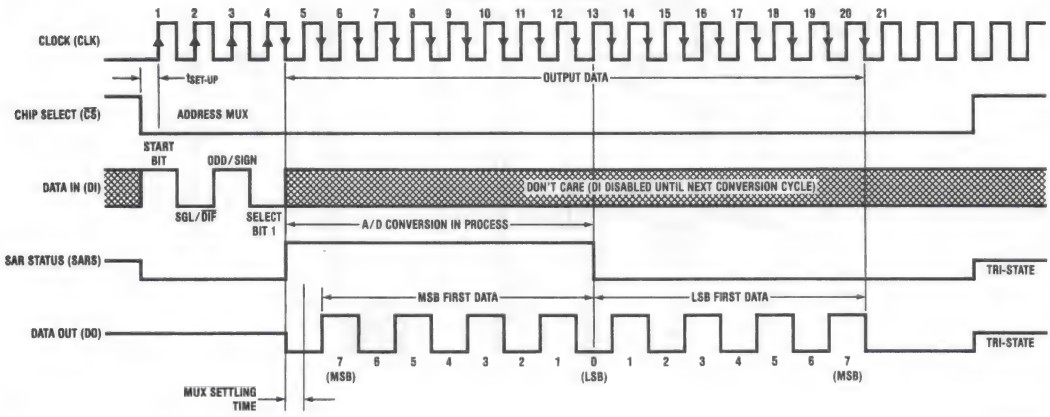
*LSB first output not available on ADC0831.

ADC0832 Timing



TL/H/5583–28

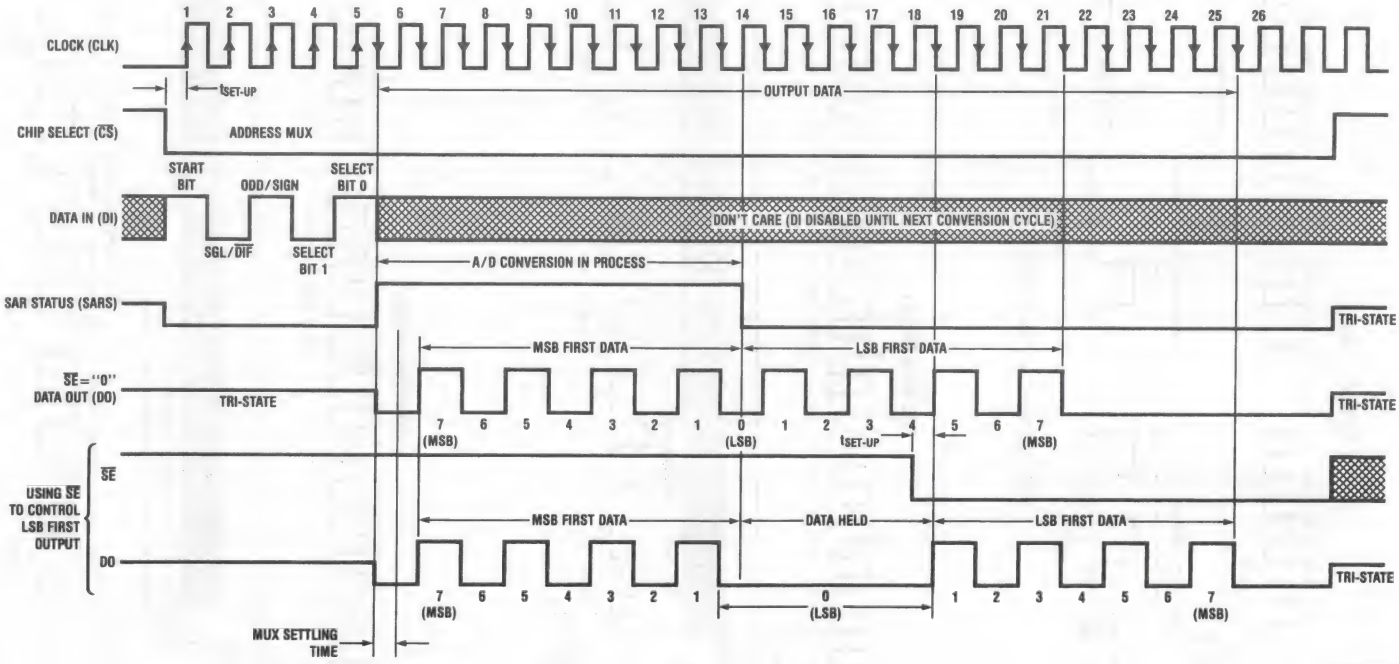
ADC0834 Timing



TL/H/5583–5

Timing Diagrams (Continued)

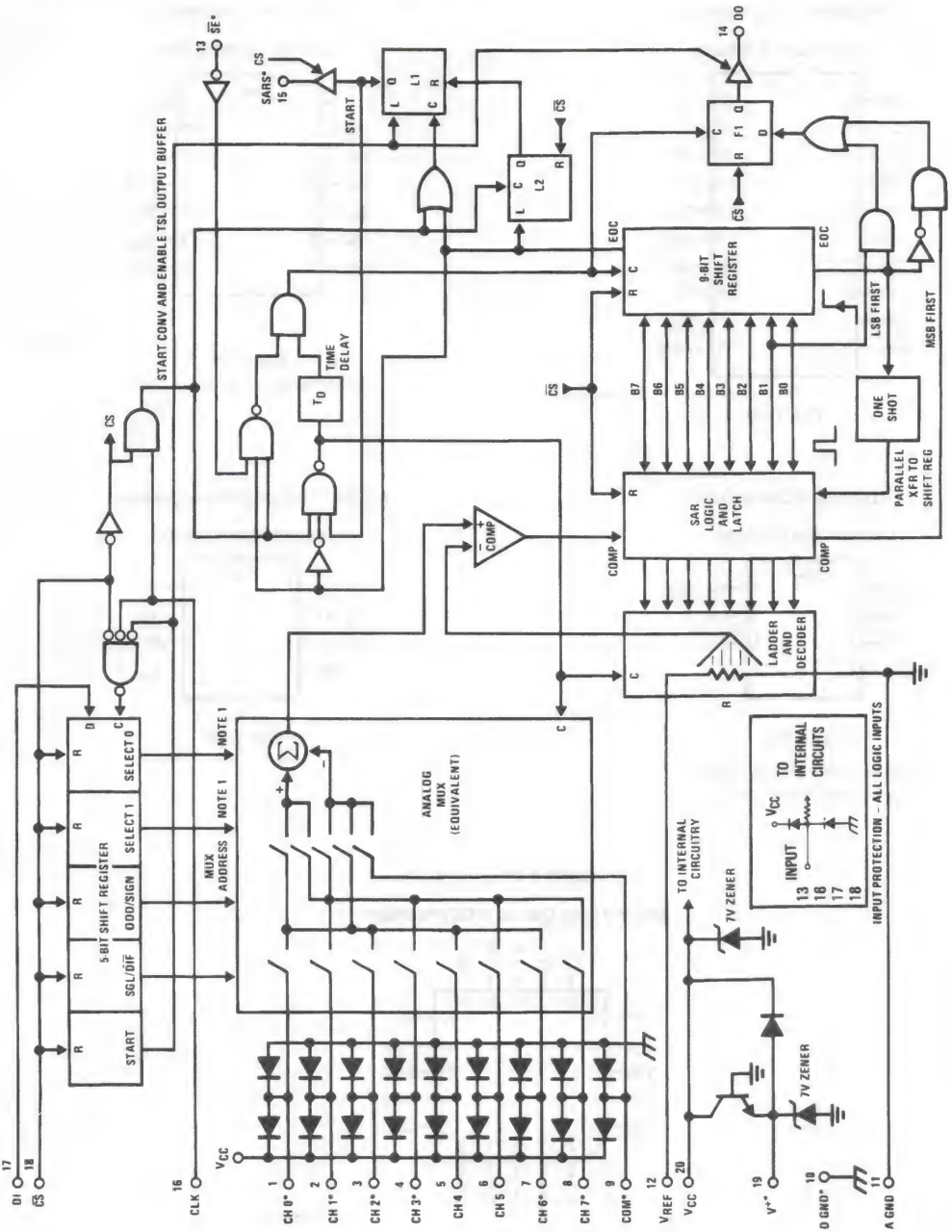
ADC0838 Timing



TL/H/5583–6

* Make sure clock edge #18 clocks in the LSB before $\overline{SE}$ is taken low

ADC0838 Functional Block Diagram



*Some of these functions/pins are not available with other options.

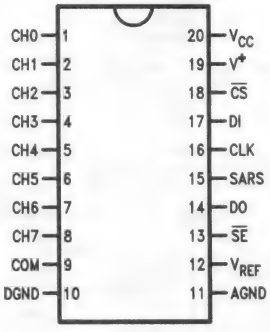
Note 1: For the ADC0834, D1 is input directly to the D input of SELECT 1. SELECT 0 is forced to a "1". For the ADC0832, DI is input directly to the DI input of ODD/SIGN. SELECT 0 is forced to a "0" and SELECT 1 is forced to a "1".

3

Connection Diagrams

ADC0838 8-Channel MUX

Dual-In-Line Package

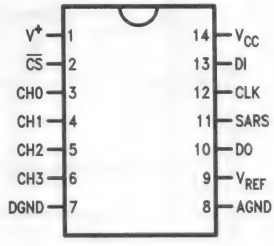


TL/H/5583–8

Top View

ADC0834 4-Channel MUX

Dual-In-Line Package



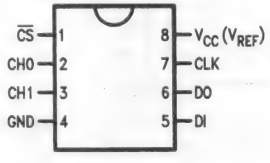
TL/H/5583–30

Top View

COM internally connected to A GND

ADC0832 2-Channel MUX

Dual-In-Line Package



TL/H/5583–31

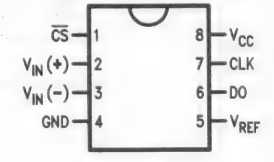
Top View

COM internally connected to GND.

V_{REF} internally connected to V_{CC}.

ADC0831 Single Differential Input

Dual-In-Line Package

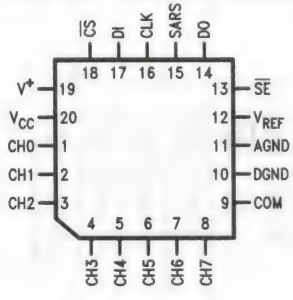


TL/H/5583–32

Top View

ADC0838 8-Channel MUX

Molded Chip Carrier (PCC) Package



TL/H/5583–33

Functional Description

1.0 MULTIPLEXER ADDRESSING

The design of these converters utilizes a sample-data comparator structure which provides for a differential analog input to be converted by a successive approximation routine.

The actual voltage converted is always the difference between an assigned "+" input terminal and a "−" input terminal. The polarity of each input terminal of the pair being converted indicates which line the converter expects to be the most positive. If the assigned "+" input is less than the "−" input the converter responds with an all zeros output code.

A unique input multiplexing scheme has been utilized to provide multiple analog channels with software-configurable single-ended, differential, or a new pseudo-differential option which will convert the difference between the voltage at any analog input and a common terminal. The analog signal conditioning required in transducer-based data acquisition systems is significantly simplified with this type of input flexibility. One converter package can now handle ground referenced inputs and true differential inputs as well as signals with some arbitrary reference voltage.

A particular input configuration is assigned during the MUX addressing sequence, prior to the start of a conversion. The MUX address selects which of the analog inputs are to be enabled and whether this input is single-ended or differential. In the differential case, it also assigns the polarity of the channels. Differential inputs are restricted to adjacent channel pairs. For example channel 0 and channel 1 may be selected as a different pair but channel 0 or 1 cannot act differentially with any other channel. In addition to selecting differential mode the sign may also be selected. Channel 0 may be selected as the positive input and channel 1 as the negative input or vice versa. This programmability is best illustrated by the MUX addressing codes shown in the following tables for the various product options.

The MUX address is shifted into the converter via the DI line. Because the ADC0831 contains only one differential input channel with a fixed polarity assignment, it does not require addressing.

The common input line on the ADC0838 can be used as a pseudo-differential input. In this mode, the voltage on this pin is treated as the "−" input for any of the other input channels. This voltage does not have to be analog ground; it can be any reference potential which is common to all of the inputs. This feature is most useful in single-supply application where the analog circuitry may be biased up to a potential other than ground and the output signals are all referred to this potential.

TABLE I. Multiplexer/Package Options

Part Number	Number of Analog Channels		Number of Package Pins
	Single-Ended	Differential	
ADC0831	1	1	8
ADC0832	2	1	8
ADC0834	4	2	14
ADC0838	8	4	20

Functional Description (Continued)

TABLE II. MUX Addressing: ADC0838

Single-Ended MUX Mode

MUX Address				Analog Single-Ended Channel #								
SGL/$\overline{\text{DIF}}$	ODD/SIGN	SELECT 1	SELECT 0	0	1	2	3	4	5	6	7	COM
1	0	0	0	+								−
1	0	0	1			+						−
1	0	1	0					+				−
1	0	1	1							+		−
1	1	0	0		+							−
1	1	0	1				+					−
1	1	1	0						+			−
1	1	1	1								+	−

Differential MUX Mode

MUX Address				Analog Differential Channel-Pair #							
SGL/$\overline{\text{DIF}}$	ODD/SIGN	SELECT		0		1		2		3	
		1	0	0	1	2	3	4	5	6	7
0	0	0	0	+	−						
0	0	0	1			+	−				
0	0	1	0					+	−		
0	0	1	1							+	−
0	1	0	0	−	+						
0	1	0	1			−	+				
0	1	1	0					−	+		
0	1	1	1							−	+

TABLE III. MUX Addressing: ADC0834

Single-Ended MUX Mode

MUX Address			Channel #			
SGL/DIF	ODD/SIGN	SELECT 1	0	1	2	3
1	0	0	+			
1	0	1			+	
1	1	0		+		
1	1	1				+

COM is internally tied to A GND

Differential MUX Mode

MUX Address			Channel #			
SGL/DIF	ODD/SIGN	SELECT 1	0	1	2	3
0	0	0	+	−		
0	0	1			+	−
0	1	0	−	+		
0	1	1			−	+

TABLE IV. MUX Addressing: ADC0832

Single-Ended MUX Mode

MUX Address		Channel #	
SGL/$\overline{\text{DIF}}$	ODD/SIGN	0	1
1	0	+	
1	1		+

COM is internally tied to A GND

Differential MUX Mode

MUX Address		Channel #	
SGL/$\overline{\text{DIF}}$	ODD/SIGN	0	1
0	0	+	−
0	1	−	+

Functional Description (Continued)

Since the input configuration is under software control, it can be modified, as required, at each conversion. A channel can be treated as a single-ended, ground referenced input for one conversion; then it can be reconfigured as part of a differential channel for another conversion. *Figure 1* illustrates the input flexibility which can be achieved.

The analog input voltages for each channel can range from 50 mV below ground to 50 mV above V_{CC} (typically 5V) without degrading conversion accuracy.

2.0 THE DIGITAL INTERFACE

A most important characteristic of these converters is their serial data link with the controlling processor. Using a serial communication format offers two very significant system improvements; it allows more function to be included in the converter package with no increase in package size and it can eliminate the transmission of low level analog signals by locating the converter right at the analog sensor; transmitting highly noise immune digital data back to the host processor.

To understand the operation of these converters it is best to refer to the Timing Diagrams and Functional Block Diagram and to follow a complete conversion sequence. For clarity a separate diagram is shown of each device.

1. A conversion is initiated by first pulling the $\overline{CS}$ (chip select) line low. This line must be held low for the entire conversion. The converter is now waiting for a start bit and its MUX assignment word.

2. A clock is then generated by the processor (if not provided continuously) and output to the A/D clock input.

3. On each rising edge of the clock the status of the data in (DI) line is clocked into the MUX address shift register. The start bit is the first logic "1" that appears on this line (all leading zeros are ignored). Following the start bit the converter expects the next 2 to 4 bits to be the MUX assignment word.

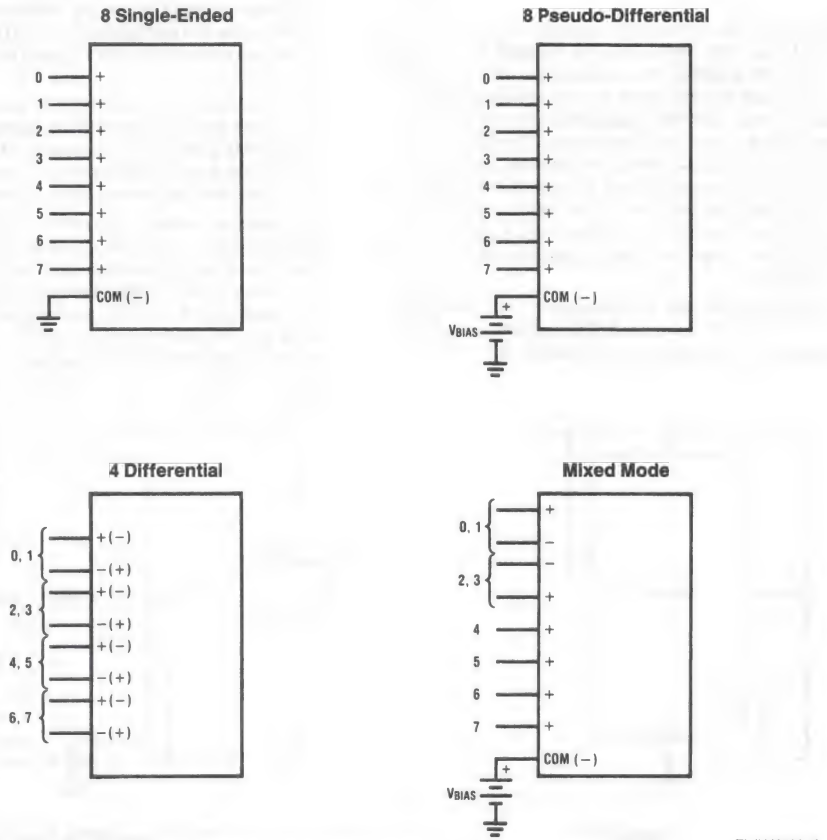


FIGURE 1. Analog Input Multiplexer Options for the ADC0838

Functional Description (Continued)

4. When the start bit has been shifted into the start location of the MUX register, the input channel has been assigned and a conversion is about to begin. An interval of ½ clock period (where nothing happens) is automatically inserted to allow the selected MUX channel to settle. The SAR status line goes high at this time to signal that a conversion is now in progress and the DI line is disabled (it no longer accepts data).

5. The data out (DO) line now comes out of TRI-STATE and provides a leading zero for this one clock period of MUX settling time.

6. When the conversion begins, the output of the SAR comparator, which indicates whether the analog input is greater than (high) or less than (low) each successive voltage from the internal resistor ladder, appears at the DO line on each falling edge of the clock. This data is the result of the conversion being shifted out (with the MSB coming first) and can be read by the processor immediately.

7. After 8 clock periods the conversion is completed. The SAR status line returns low to indicate this ½ clock cycle later.

8. If the programmer prefers, the data can be provided in an LSB first format [this makes use of the shift enable ($\overline{SE}$) control line]. All 8 bits of the result are stored in an output shift register. On devices which do not include the $\overline{SE}$ control line, the data, LSB first, is automatically shifted out the DO line, after the MSB first data stream. The DO line then goes low and stays low until $\overline{CS}$ is returned high. On the ADC0838 the $\overline{SE}$ line is brought out and if held high, the value of the LSB remains valid on the DO line. When $\overline{SE}$ is forced low, the data is then clocked out LSB first. The ADC0831 is an exception in that its data is only output in MSB first format.

9. All internal registers are cleared when the $\overline{CS}$ line is high. If another conversion is desired, $\overline{CS}$ must make a high to low transition followed by address information.

The DI and DO lines can be tied together and controlled through a bidirectional processor I/O bit with one wire. This is possible because the DI input is only "looked-at" during the MUX addressing interval while the DO line is still in a high impedance state.

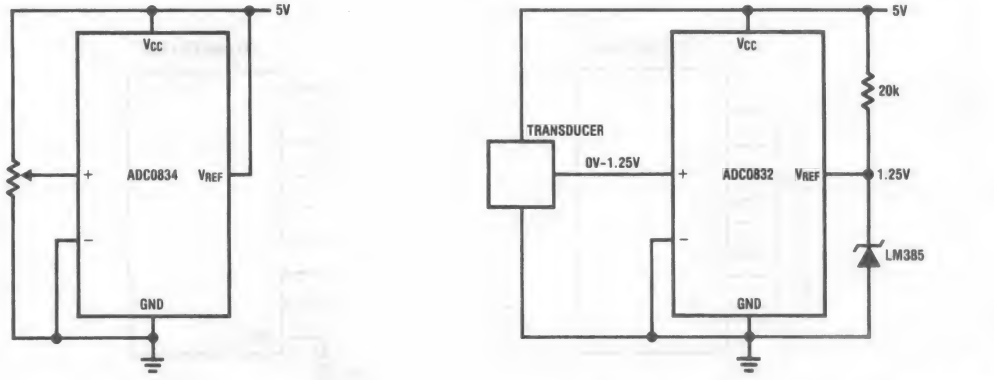
3.0 REFERENCE CONSIDERATIONS

The voltage applied to the reference input to these converters defines the voltage span of the analog input (the difference between $V_{IN(MAX)}$ and $V_{IN(MIN)}$) over which the 256 possible output codes apply. The devices can be used in either ratiometric applications or in systems requiring absolute accuracy. The reference pin must be connected to a voltage source capable of driving the reference input resistance of typically 3.5 kΩ. This pin is the top of a resistor divider string used for the successive approximation conversion.

In a ratiometric system, the analog input voltage is proportional to the voltage used for the A/D reference. This voltage is typically the system power supply, so the V_{REF} pin can be tied to V_{CC} (done internally on the ADC0832). This technique relaxes the stability requirements of the system reference as the analog input and A/D reference move together maintaining the same output code for a given input condition.

For absolute accuracy, where the analog input varies between very specific voltage limits, the reference pin can be biased with a time and temperature stable voltage source. The LM385 and LM336 reference diodes are good low current devices to use with these converters.

The maximum value of the reference is limited to the V_{CC} supply voltage. The minimum value, however, can be quite small (see Typical Performance Characteristics) to allow direct conversions of transducer outputs providing less than a 5V output span. Particular care must be taken with regard to noise pickup, circuit layout and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter (1 LSB equals V_{REF}/256).



a) Ratiometric

b) Absolute with a Reduced Span

FIGURE 2. Reference Examples

Functional Description (Continued)

4.0 THE ANALOG INPUTS

The most important feature of these converters is that they can be located right at the analog signal source and through just a few wires can communicate with a controlling processor with a highly noise immune serial bit stream. This in itself greatly minimizes circuitry to maintain analog signal accuracy which otherwise is most susceptible to noise pickup. However, a few words are in order with regard to the analog inputs should the input be noisy to begin with or possibly riding on a large common-mode voltage.

The differential input of these converters actually reduces the effects of common-mode input noise, a signal common to both selected "+" and "−" inputs for a conversion (60 Hz is most typical). The time interval between sampling the "+" input and then the "−" input is ½ of a clock period. The change in the common-mode voltage during this short time interval can cause conversion errors. For a sinusoidal common-mode signal this error is:

$$V_{error}(max) = V_{PEAK}(2\pi f_{CM}) \left(\frac{0.5}{f_{CLK}} \right)$$

where f_{CM} is the frequency of the common-mode signal,

V_{PEAK} is its peak voltage value

and f_{CLK}, is the A/D clock frequency.

For a 60 Hz common-mode signal to generate a ¼ LSB error (≈5 mV) with the converter running at 250 kHz, its peak value would have to be 6.63V which would be larger than allowed as it exceeds the maximum analog input limits.

Due to the sampling nature of the analog inputs short spikes of current enter the "+" input and exit the "−" input at the clock edges during the actual conversion. These currents decay rapidly and do not cause errors as the internal comparator is strobed at the end of a clock period. Bypass capacitors at the inputs will average these currents and cause an effective DC current to flow through the output resistance of the analog signal source. Bypass capacitors should not be used if the source resistance is greater than 1 kΩ.

This source resistance limitation is important with regard to the DC leakage currents of input multiplexer as well. The worst-case leakage current of ±1 μA over temperature will create a 1 mV input error with a 1 kΩ source resistance. An op amp RC active low pass filter can provide both impedance buffering and noise filtering should a high impedance signal source be required.

5.0 OPTIONAL ADJUSTMENTS

5.1 Zero Error

The zero of the A/D does not require adjustment. If the minimum analog input voltage value, $V_{IN(MIN)}$, is not ground a zero offset can be done. The converter can be made to output 0000 0000 digital code for this minimum input voltage by biasing any V_{IN} (−) input at this $V_{IN(MIN)}$ value. This utilizes the differential mode operation of the A/D.

The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the $V_{IN}(-)$ input and applying a small magnitude positive voltage to the $V_{IN}(+)$ input. Zero error is the difference between the actual DC input voltage which is necessary to just cause an output digital code transition from 0000 0000 to 0000 0001 and the ideal ½ LSB value (½ LSB = 9.8 mV for $V_{REF} = 5.000\ V_{DC}$).

5.2 Full-Scale

The full-scale adjustment can be made by applying a differential input voltage which is 1 ½ LSB down from the desired analog full-scale voltage range and then adjusting the magnitude of the V_{REF} input (or V_{CC} for the ADC0832) for a digital output code which is just changing from 1111 1110 to 1111 1111.

5.3 Adjusting for an Arbitrary Analog Input Voltage Range

If the analog zero voltage of the A/D is shifted away from ground (for example, to accommodate an analog input signal which does not go to ground), this new zero reference should be properly adjusted first. A V_{IN} (+) voltage which equals this desired zero reference plus ½ LSB (where the LSB is calculated for the desired analog span, using 1 LSB = analog span/256) is applied to selected "+" input and the zero reference voltage at the corresponding "−" input should then be adjusted to just obtain the 00_{HEX} to 01_{HEX} code transition.

The full-scale adjustment should be made [with the proper $V_{IN}(-)$ voltage applied] by forcing a voltage to the $V_{IN}(+)$ input which is given by:

$$V_{IN}(+) \text{ fs adj} = V_{MAX} - 1.5 \left[\frac{(V_{MAX} - V_{MIN})}{256} \right]$$

where:

V_{MAX} = the high end of the analog input range

and

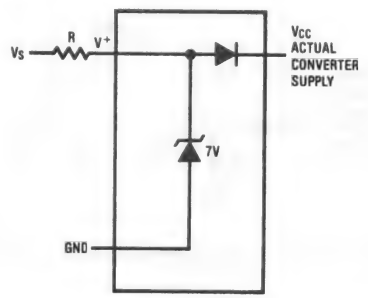
V_{MIN} = the low end (the offset zero) of the analog range.

(Both are ground referenced.)

The V_{REF} (or V_{CC}) voltage is then adjusted to provide a code change from FE_{HEX} to FF_{HEX}. This completes the adjustment procedure.

6.0 POWER SUPPLY

A unique feature of the ADC0838 and ADC0834 is the inclusion of a zener diode connected from the V^+ terminal to ground which also connects to the V_{CC} terminal (which is the actual converter supply) through a silicon diode, as shown in *Figure 3*. (See Note 3)



TL/H/5583–11

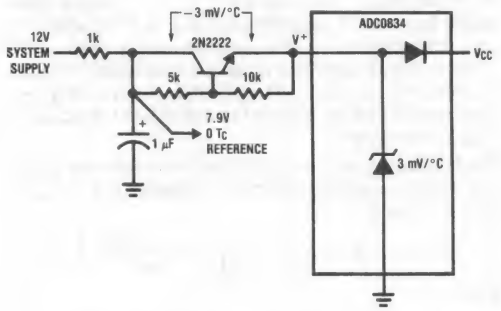
FIGURE 3. An On-Chip Shunt Regulator Diode

Functional Description (Continued)

This zener is intended for use as a shunt voltage regulator to eliminate the need for any additional regulating components. This is most desirable if the converter is to be remotely located from the system power source. *Figures 4 and 5* illustrate two useful applications of this on-board zener when an external transistor can be afforded.

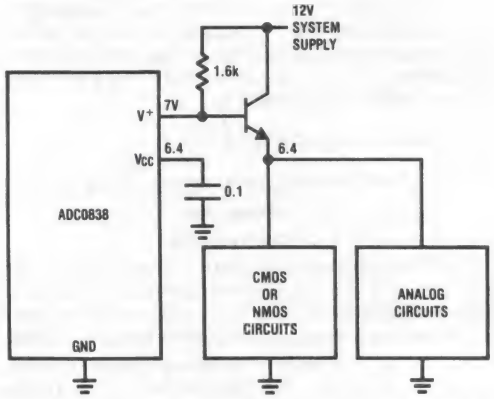
An important use of the interconnecting diode between V^+ and V_{CC} is shown in *Figures 6 and 7*. Here, this diode is used as a rectifier to allow the V_{CC} supply for the converter to be derived from the clock. The low current requirements of the A/D and the relatively high clock frequencies used (typically in the range of 10k–400 kHz) allows using the small value filter capacitor shown to keep the ripple on the V_{CC} line to well under ¼ of an LSB. The shunt zener regulator can also be used in this mode. This requires a clock voltage swing which is in excess of V_Z. A current limit for the zener is needed, either built into the clock generator or a resistor can be used from the CLK pin to the V^+ pin.

Applications



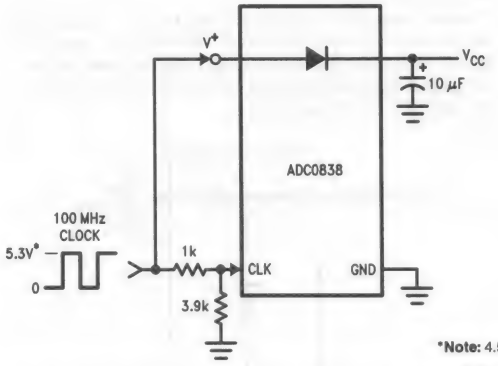
TL/H/5583–12

FIGURE 4. Operating with a Temperature Compensated Reference



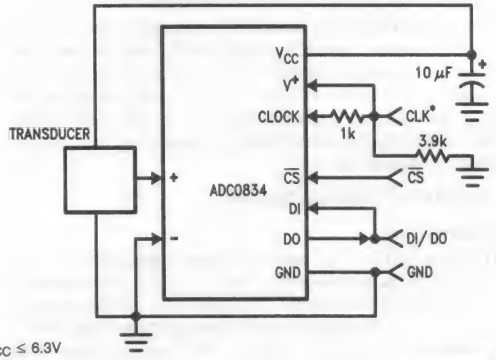
TL/H/5583–34

FIGURE 5. Using the A/D as the System Supply Regulator



TL/H/5583–35

FIGURE 6. Generating V_{CC} from the Converter Clock



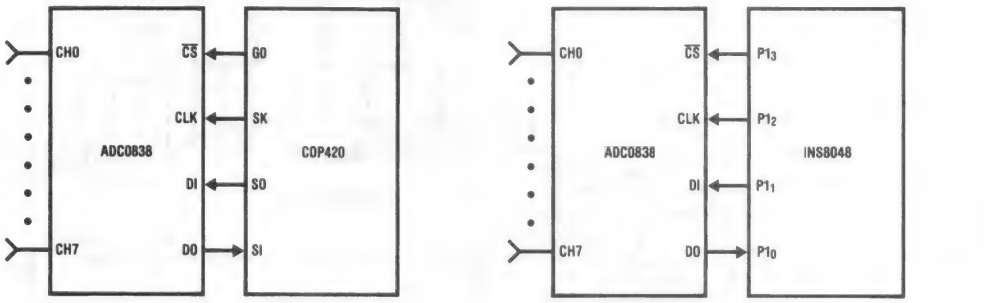
***Note:** 4.5V ≤ V_{CC} ≤ 6.3V

TL/H/5583–36

FIGURE 7. Remote Sensing—Clock and Power on 1 Wire

Applications (Continued)

Digital Link and Sample Controlling Software for the Serially Oriented COP420 and the Bit Programmable I/O INS8048



TL/H/5583–13

COP CODING EXAMPLE

Mnemonic	Instruction
LEI	ENABLES SIO's INPUT AND OUTPUT
SC	C = 1
OGI	G0 = 0 ($\overline{CS}$ = 0)
CLR A	CLEARS ACCUMULATOR
AISC 1	LOADS ACCUMULATOR WITH 1
XAS	EXCHANGES SIO WITH ACCUMULATOR AND STARTS SK CLOCK
LDD	LOADS MUX ADDRESS FROM RAM INTO ACCUMULATOR
NOP	—
XAS	LOADS MUX ADDRESS FROM ACCUMULATOR TO SIO REGISTER
↑ 8 INSTRUCTIONS ↓	
XAS	READS HIGH ORDER NIBBLE (4 BITS) INTO ACCUMULATOR
XIS	PUTS HIGH ORDER NIBBLE INTO RAM
CLR A	CLEARS ACCUMULATOR
RC	C = 0
XAS	READS LOW ORDER NIBBLE INTO ACCUMULATOR AND STOPS SK
XIS	PUTS LOW ORDER NIBBLE INTO RAM
OGI	G0 = 1 ($\overline{CS}$ = 1)
LEI	DISABLES SIO's INPUT AND OUTPUT

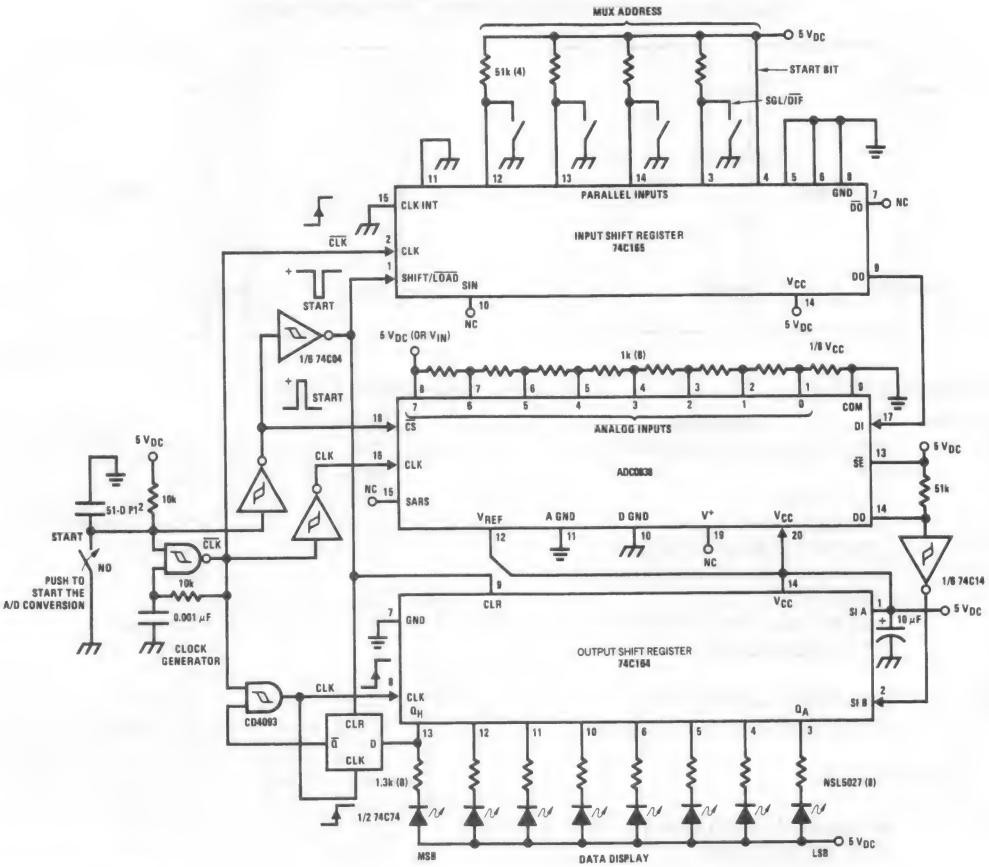
8048 CODING EXAMPLE

	Mnemonic		Instruction
START:	ANL	P1, #0F7H	;SELECT A/D ($\overline{CS}$ = 0)
	MOV	B, #5	;BIT COUNTER ← 5
	MOV	A, #ADDR	;A ← MUX ADDRESS
LOOP 1:	RRC	A	;CY ← ADDRESS BIT
	JC	ONE	;TEST BIT
			;BIT = 0
ZERO:	ANL	P1, #0FEH	;DI ← 0
	JMP	CONT	;CONTINUE
			;BIT = 1
ONE:	ORL	P1, #1	;DI ← 1
CONT:	CALL	PULSE	;PULSE SK 0 → 1 → 0
	DJNZ	B, LOOP 1	;CONTINUE UNTIL DONE
	CALL	PULSE	;EXTRA CLOCK FOR SYNC
	MOV	B, #8	;BIT COUNTER ← 8
LOOP 2:	CALL	PULSE	;PULSE SK 0 → 1 → 0
	IN	A, P1	;CY ← DO
	RRC	A	
	RRC	A	
	MOV	A, C	;A ← RESULT
	RLC	A	;A(0) ← BIT AND SHIFT
	MOV	C, A	;C ← RESULT
	DJNZ	B, LOOP 2	;CONTINUE UNTIL DONE
RETR			
			;PULSE SUBROUTINE
PULSE:	ORL	P1, #04	;SK ← 1
	NOP		;DELAY
	ANL	P1, #0FBH	;SK ← 0
	RET		

3

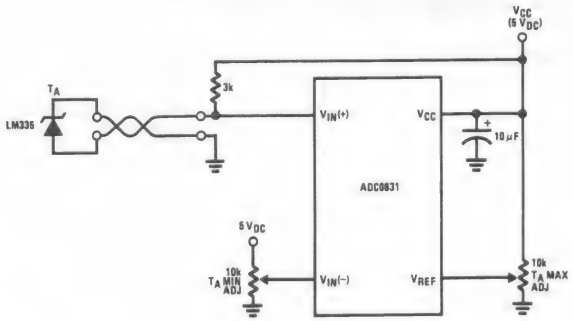
Applications (Continued)

A "Stand-Alone" Hook-Up for ADC0838 Evaluation



*Pinouts shown for ADC0838.
For all other products tie to pin functions as shown.

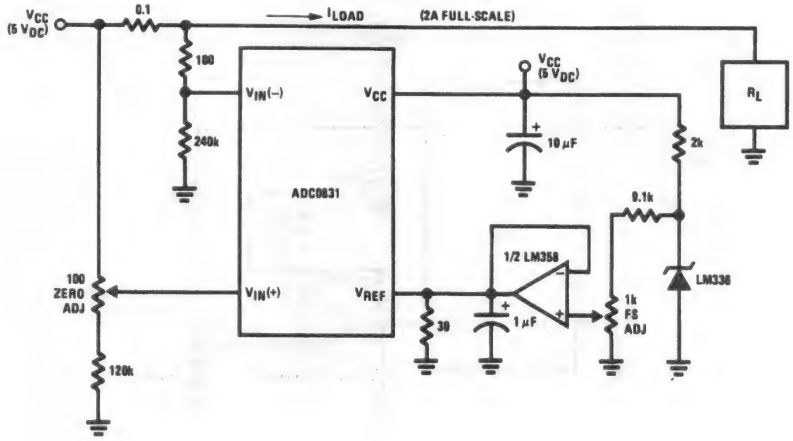
Low-Cost Remote Temperature Sensor



TL/H/5583–14

Applications (Continued)

Digitizing a Current Flow



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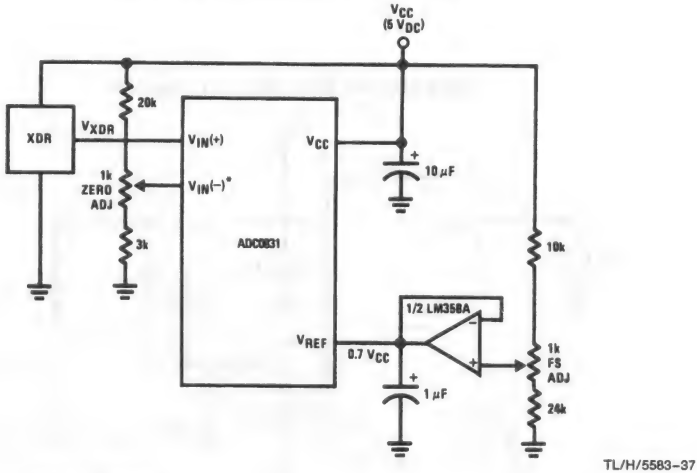
Operating with Ratiometric Transducers



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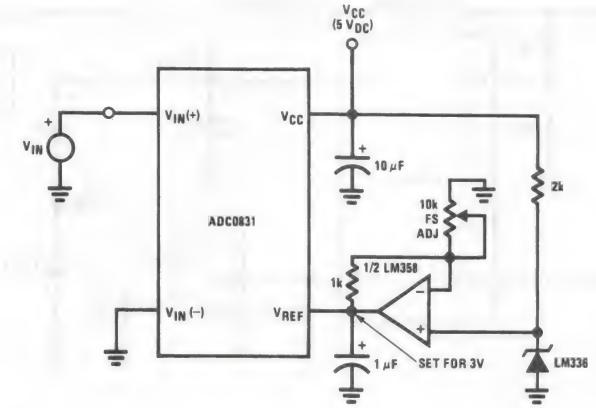
*$V_{IN}(-)$ = 0.15 V_{CC}
15% of $V_{CC} \leq V_{XDR} \leq$ 85% of V_{CC}

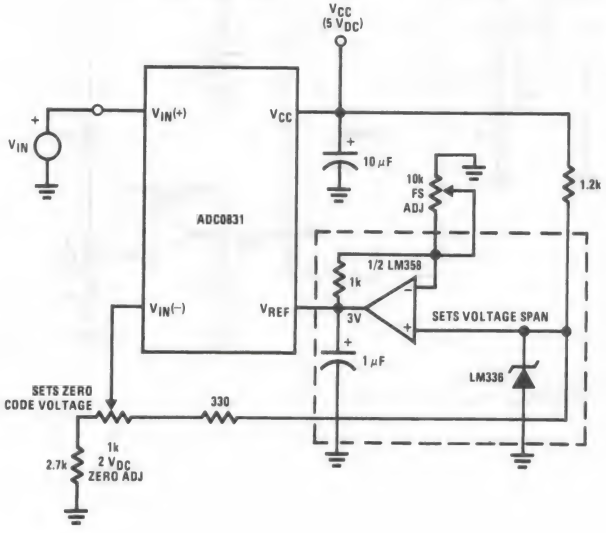
3

Applications (Continued)

Span Adjust: $0V \leq V_{IN} \leq 3V$



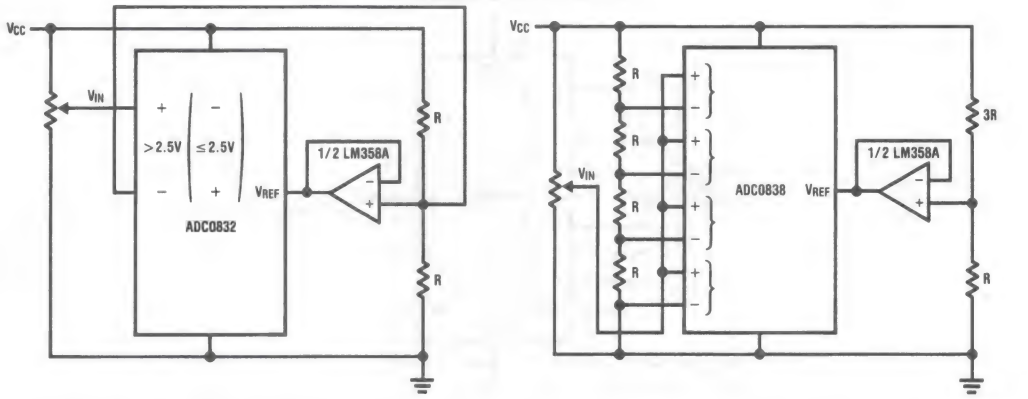
Zero-Shift and Span Adjust: $2V \leq V_{IN} \leq 5V$



TL/H/5583–16

Applications (Continued)

Obtaining Higher Resolution



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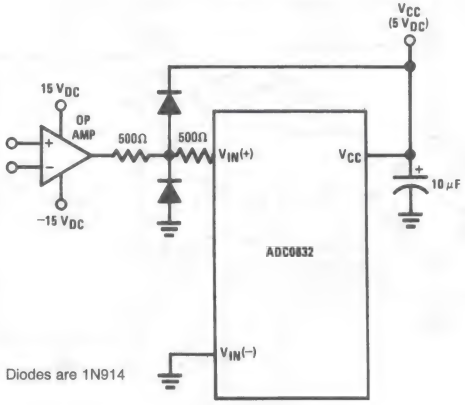
Controller performs a routine to determine which input polarity (9-bit example) or which channel pair (10-bit example) provides a non-zero output code. This information provides the extra bits.

a) 9-Bit A/D **b)10-Bit A/D**

Protecting the Input



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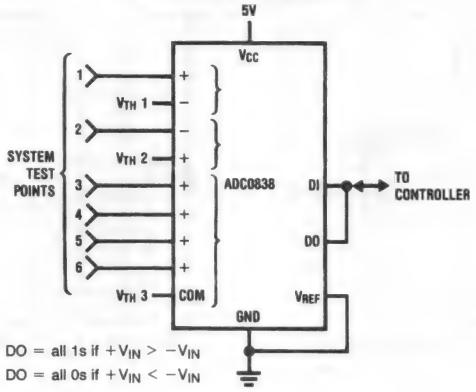
Applications (Continued)

High Accuracy Comparators



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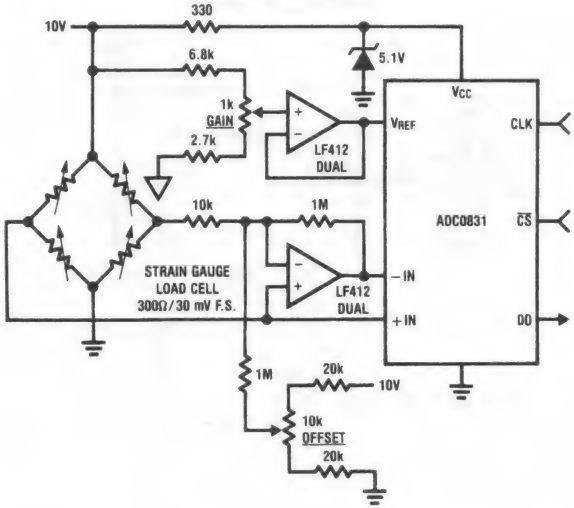
Digital Load Cell



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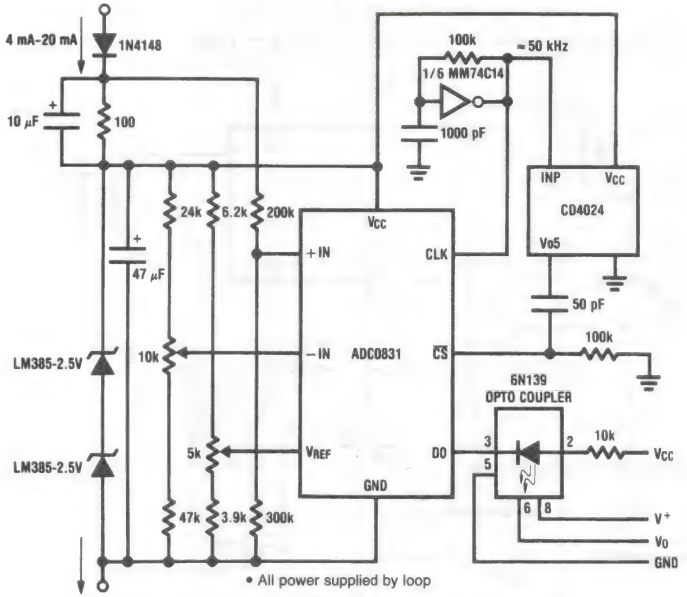
- Uses one more wire than load cell itself
- Two mini-DIPs could be mounted inside load cell for digital output transducer
- Electronic offset and gain trims relax mechanical specs for gauge factor and offset
- Low level cell output is converted immediately for high noise immunity

Applications (Continued)

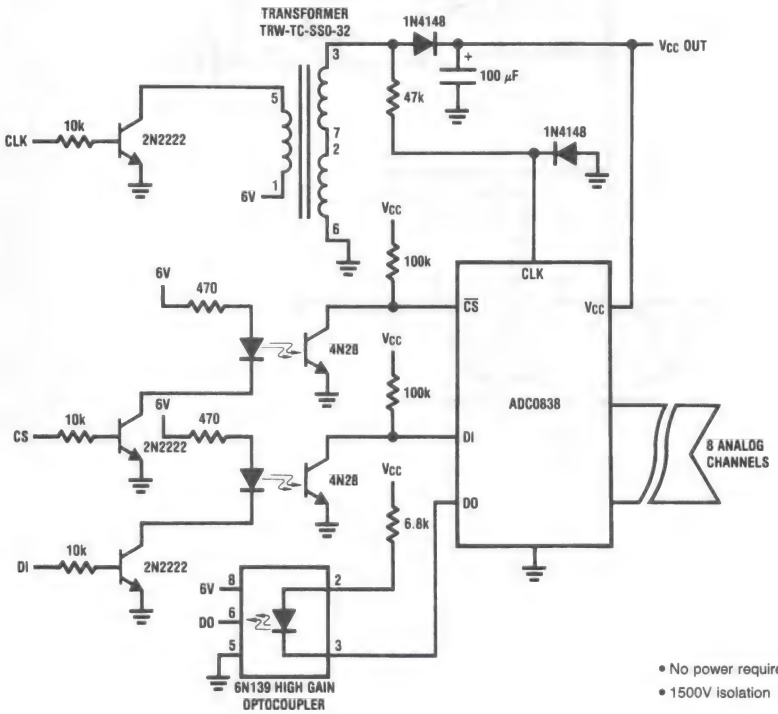
4 mA–20 mA Current Loop Converter



- All power supplied by loop
- 1500V isolation at output

TL/H/5583–20

Isolated Data Converter



- No power required remotely
- 1500V isolation

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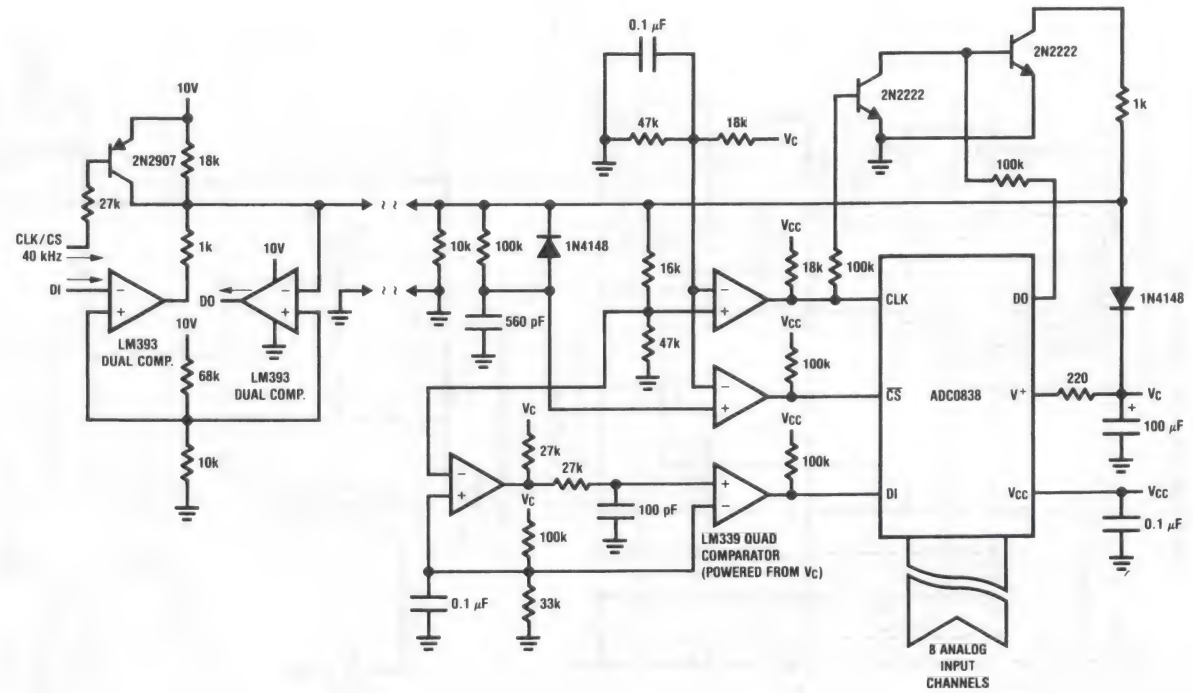
Applications (Continued)

Two Wire Interface for 8 Channels



- No additional connections
- $\overline{CS}$ derived from extended high on CLK line > 100 μs
- Timing arranged for 40 kHz, could be changed up or down by component change
- 10% CLK frequency change without component change OK

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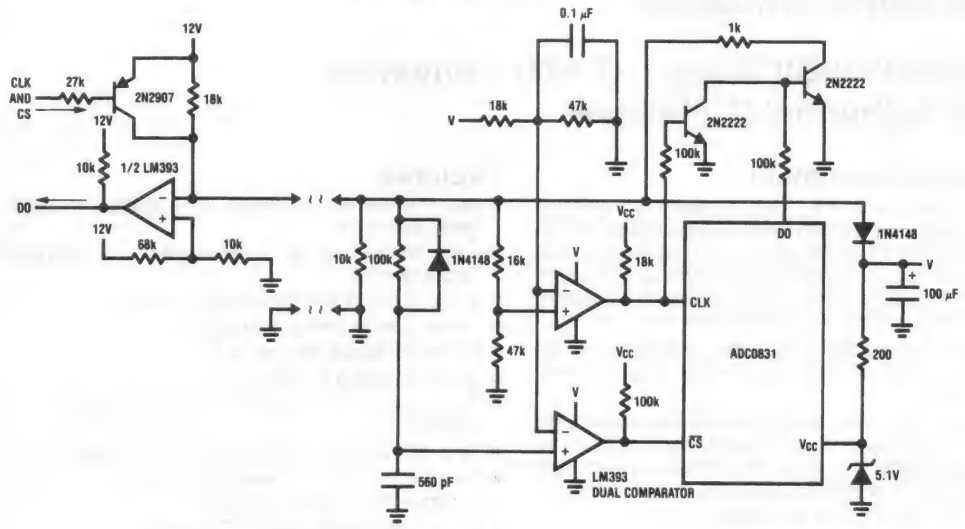
Applications (Continued)

Two Wire 1-Channel Interface



- Simpler version of 8-channel
- $\overline{CS}$ derived from long CLK pulse

TL/H/5583–22

Ordering Information

Part Number	Analog Input Channels	Total Unadjusted Error	Package	Temperature Range
ADC0831BJ	1	±½	Hermetic (J)	−55°C to +125°C
ADC0831BCJ			Hermetic (J)	−40°C to +85°C
ADC0831BCN			Molded (N)	0°C to +70°C
ADC0831CCJ		±1	Hermetic (J)	−40°C to +85°C
ADC0831CCN			Molded (N)	0°C to +70°C
ADC0832BJ	2	±½	Hermetic (J)	−55°C to +125°C
ADC0832BCJ			Hermetic (J)	−40°C to +85°C
ADC0832BCN			Molded (N)	0°C to +70°C
ADC0832CCJ		±1	Hermetic (J)	−40°C to +85°C
ADC0832CCN			Molded (N)	0°C to +70°C
ADC0834BJ	4	±½	Hermetic (J)	−55°C to +125°C
ADC0834BCJ			Hermetic (J)	−40°C to +85°C
ADC0834BCN			Molded (N)	0°C to +70°C
ADC0834CCJ		±1	Hermetic (J)	−40°C to +85°C
ADC0834CCN			Molded (N)	0°C to +70°C
ADC0838BJ	8	±½	Hermetic (J)	−55°C to +125°C
ADC0838BCJ			Hermetic (J)	−40°C to +85°C
ADC0838BCV			PCC (V)	0°C to +70°C
ADC0838BCN			Molded (N)	0°C to +70°C
ADC0838CCJ		±1	Hermetic (J)	−40°C to +85°C
ADC0838CCV			PCC (V)	0°C to +70°C
ADC0838CCN			Molded (N)	0°C to +70°C

See NS Package Number J08A, J14A, J20A, N08E, N14A, N20A or V20A

National Semiconductor

ADC0833 8-Bit Serial I/O A/D Converter with 4-Channel Multiplexer

General Description

The ADC0833 series is an 8-bit successive approximation A/D converter with a serial I/O and configurable input multiplexer with 4 channels. The serial I/O is configured to comply with the NSC MICROWIRE™ serial data exchange standard for easy interface to the COPS™ family of processors, as well as with standard shift registers or μPs.

The 4-channel multiplexer is software configured for single-ended or differential inputs when channel assigned by a 4-bit serial word.

The differential analog voltage input allows increasing the common-mode rejection and offsetting the analog zero input voltage value. In addition, the voltage reference input can be adjusted to allow encoding any smaller analog voltage span to the full 8 bits of resolution.

Key Specifications

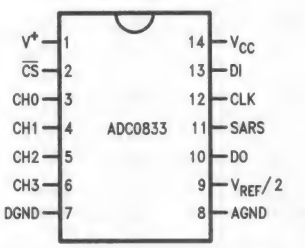
- Resolution — 8 Bits
- Total Unadjusted Error — ± ½ LSB and ± 1 LSB
- Single Supply — 5 V_{DC}
- Low Power — 23 mW
- Conversion Time — 32 μs

Features

- NSC MICROWIRE compatible—direct interface to COPS family processors
- Easy interface to all microprocessors, or operates "stand alone"
- Works with 2.5V (LM336) voltage reference
- No full-scale or zero adjust required
- Differential analog voltage inputs
- 4-channel analog multiplexer
- Shunt regulator allows operation with high voltage supplies
- 0V to 5V input range with single 5V power supply
- Remote operation with serial digital data link
- TTL/MOS input/output compatible
- 0.3" standard width 14-pin DIP package

Connection and Functional Diagrams

Dual-In-Line Package



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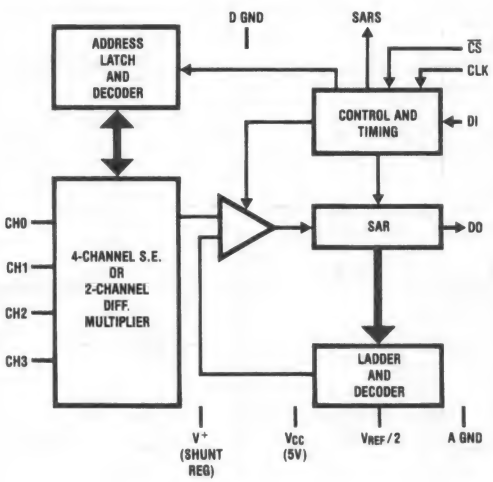
Top View

Order Number ADC0833BCJ, ADC0833BJ, ADC0833CJ, ADC0833CCJ, ADC0833BCN or ADC0833CCN
See NS Package Number J14A or N14A



TL/H/5607–1

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Current into V+ (Note 3)	15 mA
Supply Voltage, V_{CC} (Note 3)	6.5V
Voltage	
Logic Inputs	−0.3V to V_{CC} + 0.3V
Analog Inputs	−0.3V to V_{CC} + 0.3V
Input Current per Pin (Note 4)	±5 mA
Package Input Current (Note 4)	±20 mA
Storage Temperature	−65°C to + 150°C
Package Dissipation at T_A = 25°C (Board Mount)	0.8W
Lead Temperature (Soldering, 10 sec.)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
ESD Susceptibility (Note 5)	2000V

Operating Conditions (Notes 1 & 2)

Supply Voltage, V_{CC}	4.5 V_{DC} to 6.3 V_{DC}
Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0833BJ, ADC0833CJ	$-55°C \le T_A \le 125°C$
ADC0833BCJ, ADC0833CCJ	$-40°C \le T_A \le 85°C$
ADC0833BCN, ADC0833CCN	$0°C \le T_A \le 70°C$

Electrical Characteristics

The following specifications apply for $V_{CC} = V^+ = 5V$, f_{CLK} = 250 kHz and $V_{REF}/2 \le (V_{CC} + 0.1V)$ unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_j = 25°C$.

Parameter	Conditions	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
CONVERTER AND MULTIPLEXER CHARACTERISTICS					
Total Unadjusted Error	$V_{REF}/2$ Forced to 2.500 V_{DC}				
ADC0833BCN			±½	**±½**	LSB
ADC0883BJ, BCJ			**±½**		LSB
ADC0833CCN			±1	**±1**	LSB
ADC0833CJ, CCJ			**±1**		LSB
Minimum Total Ladder Resistance (Note 9)					
ADC0833BCJ/CCJ/BJ/CJ		7.0	**2.6**		kΩ
ADC0833BCN/CCN		7.0	2.6	**2.6**	kΩ
Maximum Total Ladder Resistance (Note 9)					
ADC0833BCJ/CCJ/BJ/CJ		7.0	**11.8**		kΩ
ADC0833BCN/CCN		7.0	10.8	**11.8**	kΩ
Minimum Common-Mode Input Range (Note 10)	All MUX Inputs and COM Input				
ADC0833BCJ/CCJ/BJ/CJ			**GND−0.05**		V
ADC0833BCN/CCN			GND−0.05	**GND−0.05**	V
Maximum Common-Mode Input Range (Note 10)	All MUX Inputs and COM Input				
ADC0833BCJ/CCJ/BJ/CJ			**V_{CC}+0.05**		V
ADC0833BCN/CCN			V_{CC}+0.05	**V_{CC}+0.05**	V
DC Common-Mode Error					
ADC0833BCJ/CCJ/BJ/CJ		±1/16	±¼		LSB
ADC0833BCN/CCN		±1/16	±¼	±¼	LSB
Change In Zero Error From V_{CC}=5V To Internal Zener Operation (Note 3)	15mA Into V+ V_{CC}=N.C. $V_{REF}/2$ = 2.500V				
ADC0833BCJ/CCJ/BJ/CJ			**1**		LSB
ADC0833BCN/CCN			1	**1**	LSB

Electrical Characteristics The following specifications apply for $V_{CC} = V^+ = 5V$, $f_{CLK} = 250$ kHz and $V_{REF}/2 \leq (V_{CC} + 0.1V)$ unless otherwise specified. **Boldface limits apply from t_{MIN} to t_{MAX};** all other limits $T_A = T_j = 25°C$. (Continued)

Parameter	Conditions	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
CONVERTER AND MULTIPLEXER CHARACTERISTICS (Continued)					
V_Z, Minimum Internal Diode Breakdown (At V+) (Note 3)	15mA Into V+				
ADC0833BCJ/CCJ/BJ/CJ			**6.3**		V
ADC0833BCN/CCN			6.3	**6.3**	V
V_Z, Maximum Internal Diode Breakdown (At V+) (Note 3)	15mA Into V+				
ADC0833BCJ/CCJ/BJ/CJ			**8.5**		V
ADC0833BCN/CCN			8.5	**8.5**	V
Power Supply Sensitivity	$V_{CC} = 5V \pm 5\%$				
ADC0833BCJ/CCJ/BJ/CJ		$\pm 1/16$	$\pm 1/8$		LSB
ADC0833BCN/CCN		$\pm 1/16$	$\pm 1/4$	$\pm 1/4$	LSB
I_{OFF}, Off Channel Leakage Current (Note 11)	On Channel = 5V, Off Channel = 0V				
ADC0833BCJ/CCJ/BJ/CJ			**−1**		μA
			−200		nA
ADC0833BCN/CCN				**−1**	μA
			−200		nA
	On Channel = 0V, Off Channel = 5V				
ADC0833BCJ/CCJ/BJ/CJ			**1**		μA
			200		nA
ADC0833BCN/CCN				**1**	μA
			200		nA
I_{ON}, On Channel Leakage Current (Note 11)	On Channel = 5V, Off Channel = 0V				
ADC083BCJ/CCJ/BJ/CJ			**1**		μA
			200		nA
ADC0833BCN/CCN				**1**	μA
			200		nA
	On Channel = 0V, Off Channel = 5V				
ADC083BCJ/CCJ/BJ/CJ			**−1**		μA
			−200		nA
ADC0833BCN/CCN				**−1**	μA
			−200		nA
DIGITAL AND DC CHARACTERISTICS					
$V_{IN(1)}$, Logical "1" Input Voltage	$V_{CC} = 5.25V$				
ADC0833BCJ/CCJ/BJ/CJ			**2.0**		V
ADC0833BCN/CCN			2.0	**2.0**	V
$V_{IN(0)}$, Logical "0" Input Voltage	$V_{CC} = 4.75V$				
ADC0833BCJ/CCJ/BJ/CJ			**0.8**		V
ADC0833BCN/CCN			0.8	**0.8**	V
$I_{IN(1)}$, Logical "1" Input Current	$V_{IN} = V_{CC}$				
ADC0833BCJ/CCJ/BJ/CJ		0.005	**1**		μA
ADC0833BCN/CCN		0.005	1	**1**	μA

Electrical Characteristics The following specifications apply for $V_{CC} = V^+ = 5V$, $f_{CLK} = 250$ kHz and $V_{REF}/2 \leq (V_{CC} + 0.1V)$ unless otherwise specified. **Boldface limits apply from t_{MIN} to t_{MAX};** all other limits $T_A = T_j = 25°C$. (Continued)

Parameter	Conditions	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
DIGITAL AND DC CHARACTERISTICS (Continued)					
$I_{IN(0)}$, Logical "0" Input Current	$V_{IN} = 0V$				
ADC0833BCJ/CCJ/BJ/CJ		−0.005	**−1**		μA
ADC0833BCN/CCN		−0.005	−1	**−1**	μA
$V_{OUT(1)}$, Logical "1" Output Voltage	$V_{CC} = 4.75V$				
ADC0833BCJ/CCJ/BJ/CJ	$I_{OUT} = -360\mu A$		**2.4**		V
ADC0833BCN/CCN			2.4	**2.4**	V
ADC0833BCJ/CCJ/BJ/CJ	$I_{OUT} = -10\mu A$		**4.5**		V
ADC0833BCN/CCN			4.5	**4.5**	V
$V_{OUT(0)}$, Logical "0" Output Voltage	$I_{OUT} = 1.6mA$, $V_{CC} = 4.75V$				
ADC0833BCJ/CCJ/BJ/CJ			**0.4**		V
ADC0833BCN/CCN			0.4	**0.4**	V
I_{OUT}, TRI-STATE Output Current (DO, SARS)					
ADC0833BCJ/CCJ/BJ/CJ	$V_{OUT} = 0.4V$	−0.1	**−3**		μA
ADC0833BCN/CCN		−0.1	−3	**−3**	μA
ADC0833BCJ/CCJ/BJ/CJ	$V_{OUT} = 5V$	0.1	**3**		μA
ADC0833BCN/CCN		0.1	3	**3**	μA
I_{SOURCE}	V_{OUT} Short to GND				
ADC0833BCJ/CCJ/BJ/CJ		−14	**−6.5**		mA
ADC0833BCN/CCN		−14	−7.5	**−6.5**	mA
I_{SINK}	V_{OUT} Short to V_{CC}				
ADC0833BCJ/CCJ/BJ/CJ		16	**8.0**		mA
ADC0833BCN/CCN		16	9.0	**8.0**	mA
I_{CC}, Supply Current (Note 3)	$V_{REF}/2$ Open Circuit				
ADC0833BCJ/CCJ/BJ/CJ		0.9	**4.5**		mA
ADC0833BCN/CCN		0.9	4.5	**4.5**	mA

AC Electrical Characteristics

The following specifications apply for $V_{CC} = V^+ = 5V$ and $t_r = t_f = 20$ ns unless otherwise specified. These limits apply for $T_A = T_j = 25°C$.

Parameter	Conditions	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
f_{CLK}, Clock Frequency Min Max			10	 400	kHz kHz
T_C, Conversion Time	Not including MUX Addressing Time		8		$1/f_{CLK}$
Clock Duty Cycle (Note 12) Min Max				40 60	% %
t_{SET-UP}, $\overline{CS}$ Falling Edge or Data Input Valid to CLK Rising Edge				250	ns
t_{HOLD}, Data Input Valid after CLK Rising Edge				90	ns
t_{pd1}, t_{pd0}—CLK Falling Edge to Output Data Valid (Note 13)	$C_L = 100$ pF Data MSB First Data LSB First	 650 250		 1500 600	 ns ns
t_{1H}, t_{0H}—Rising Edge of $\overline{CS}$ to Data Output and SARS Hi-Z	$C_L = 10$ pF, $R_L = 10k$ $C_L = 100$ pF, $R_L = 2k$ (see TRI-STATE Test Circuits)	125	 500	250 500	ns ns
C_{IN}, Capacitance of Logic Input		5			pF
C_{OUT}, Capacitance of Logic Outputs		5			pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to the ground pins.

Note 3: Internal zener diodes (approx. 7V) are connected from V^+ to GND and V_{CC} to GND. The zener at V^+ can operate as a shunt regulator and is connected to V_{CC} via a conventional diode. Since the zener voltage equals the A/D's breakdown voltage, the diode insures that V_{CC} will be below breakdown when the device is powered from V^+. Functionality is therefore guaranteed for V^+ operation even though the resultant voltage at V_{CC} may exceed the specified Absolute Max. of 6.5V. It is recommended that a resistor be used to limit the max. current into V^+.

Note 4: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 5 mA or less. The 20 mA package input current limits the number of pins that can exceed the power supply boundaries with a 5 mA current limit to four.

Note 5: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 6: Typicals are at 25°C and represent most likely parametric norm.

Note 7: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 8: Design limits are guaranteed but not 100% tested. These limits are not used to calculate outgoing quality levels.

Note 9: See Applications, section 3.0.

Note 10: For $V_{IN}(-) \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input (see Block Diagram) which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct—especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

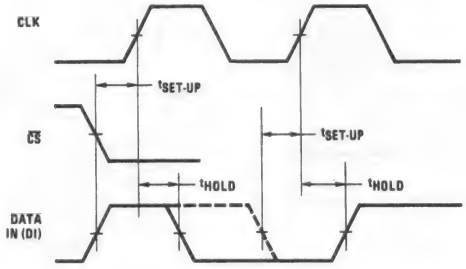
Note 11: Leakage current is measured with the clock not switching.

Note 12: A 40% to 60% clock duty cycle range insures proper operation at all clock frequencies. In the case that an available clock has a duty cycle outside of these limits, the minimum time the clock is high or the minimum time the clock is low must be at least 1 μs. The maximum time the clock can be high is 60 μs. The clocked can be stopped when low so long as the analog input voltage remains stable.

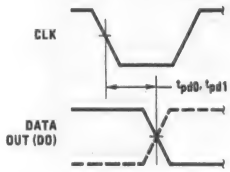
Note 13: Since data, MSB first, is the output of the comparator used in the successive approximation loop, an additional delay is built in (see Block Diagram) to allow for comparator response time.

Timing Diagrams

Data Input Timing

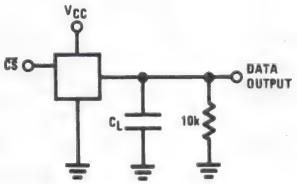


Data Output Timing

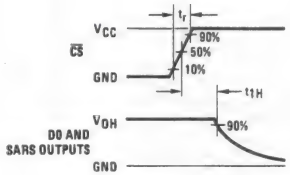


TRI-STATE Test Circuits and Waveforms

t_{1H}

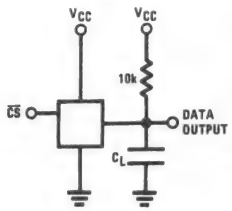


t_{1H}, $C_L = 10$ pF

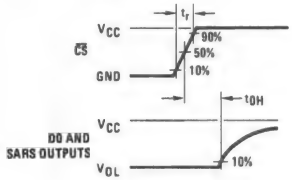


$t_r = 20$ ns

t_{0H}

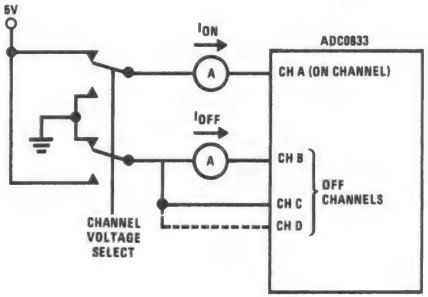


t_{0H}, $C_L = 10$ pF



$t_r = 20$ ns

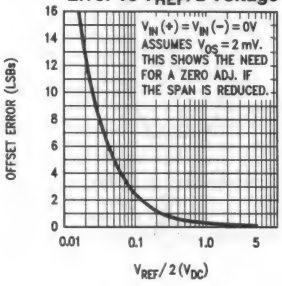
Leakage Current Test Circuit



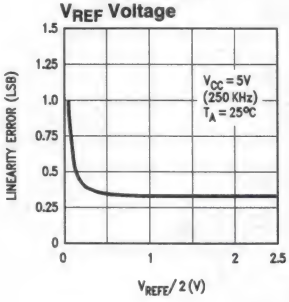
TL/H/5607–2

Typical Performance Characteristics

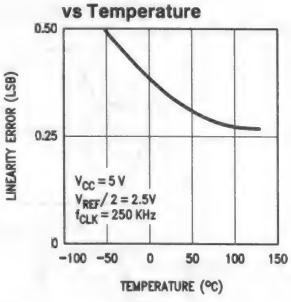
Effect of Unadjusted Offset Error vs $V_{REF}/2$ Voltage



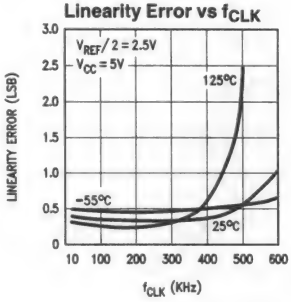
Linearity Error vs V_{REF} Voltage



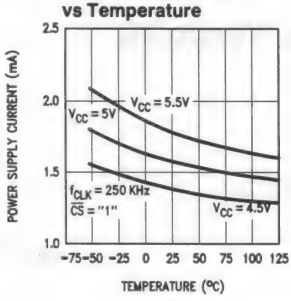
Linearity Error vs Temperature



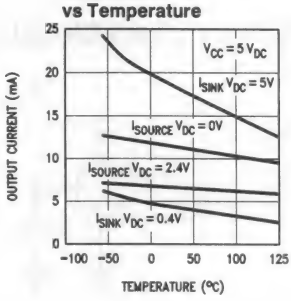
Linearity Error vs f_{CLK}



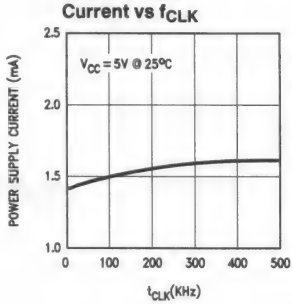
Power Supply Current vs Temperature



Output Current vs Temperature



Power Supply Current vs f_{CLK}



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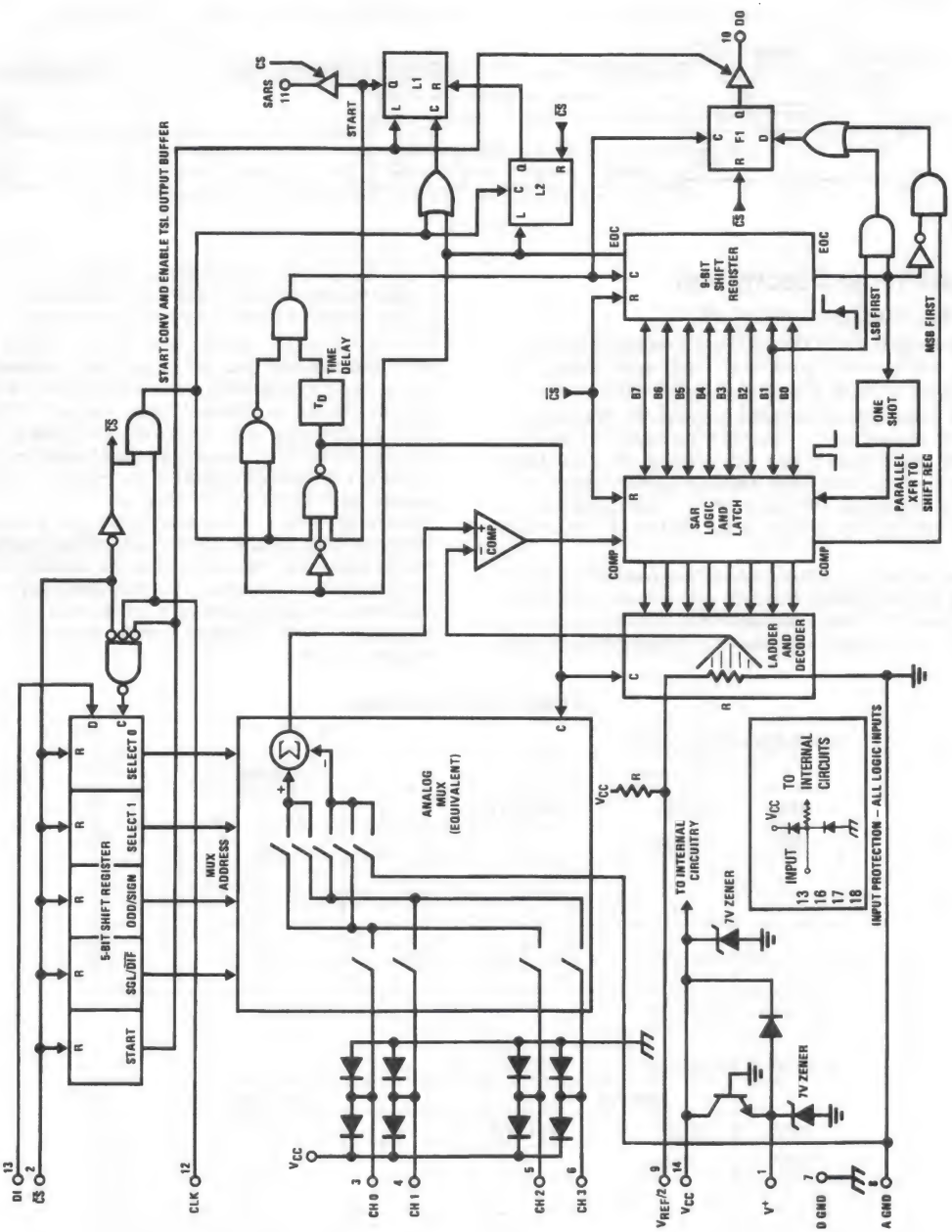
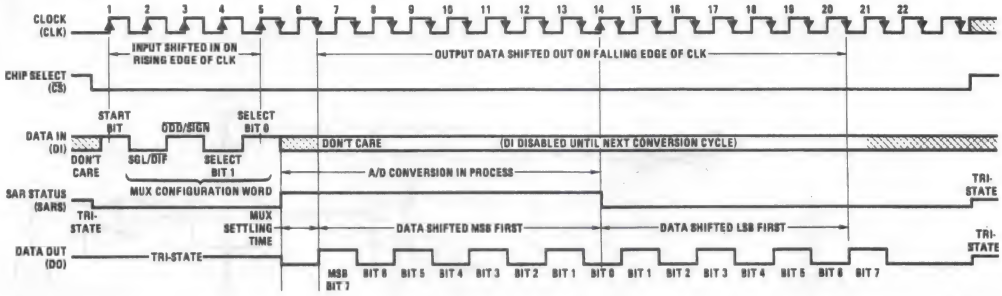
ADC0833 Functional Block Diagram



Timing Diagram


CLOCK (CLK)
1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20 21 22
INPUT SHIFTED IN ON RISING EDGE OF CLK
OUTPUT DATA SHIFTED OUT ON FALLING EDGE OF CLK
CHIP SELECT (CS)
DATA IN (DI)
START BIT
ODD/SIGN
SELECT BIT 0
DON'T CARE
SGL/DIF
SELECT BIT 1
DON'T CARE
(DI DISABLED UNTIL NEXT CONVERSION CYCLE)
A/D CONVERSION IN PROCESS
MUX CONFIGURATION WORD
SAR STATUS (SARS)
TRI-STATE
MUX SETTLING TIME
DATA SHIFTED MSB FIRST
DATA SHIFTED LSB FIRST
TRI-STATE
DATA OUT (DO)
TRI-STATE
MSB BIT 7
BIT 6 BIT 5 BIT 4 BIT 3 BIT 2 BIT 1 BIT 0 BIT 1 BIT 2 BIT 3 BIT 4 BIT 5 BIT 6 BIT 7

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Functional Description

1.0 MULTIPLEXER ADDRESSING

The design of the ADC0833 utilizes a sample-data comparator structure which provides for a differential analog input to be converted by a successive approximation routine.

The actual voltage converted is always the difference between an assigned "+" input terminal and a "−" input terminal. The polarity of each input terminal of the pair being converted indicates which line the converter expects to be the most positive. If the assigned "+" input is less than the "−" input the converter responds with an all zeros output code.

A unique input multiplexing scheme has been utilized to provide multiple analog channels with software-configurable single-ended (ground referred) or differential inputs. The analog signal conditioning required in transducer-based data acquisition systems is significantly simplified with this type of input flexibility. One converter package can now handle ground referenced inputs and true differential inputs.

A particular input configuration is assigned during the MUX addressing sequence, prior to the start of a conversion. The MUX address selects which of the analog inputs are to be enabled and whether this input is single-ended or differential. In the differential case, it also assigns the polarity of the channels. Differential inputs are restricted to adjacent channel pairs. For example channel 0 and channel 1 may be selected as a differential pair. Channel 0 or 1 cannot act differentially with any other channel. In addition to selecting differential mode the sign may also be selected. Channel 0 may be selected as the positive input and channel 1 as the negative input or vice versa. This programmability is best illustrated by the MUX addressing codes shown in the following table. The MUX address is shifted into the converter through the DI line.

TABLE I. MUX Addressing

Single-Ended MUX Mode

Address				Channel #			
SGL/ $\overline{\text{DIF}}$	$\overline{\text{ODD}}$/ $\overline{\text{SIGN}}$	SELECT 1	SELECT 0	0	1	2	3
1	0	0	1	+			
1	0	1	1			+	
1	1	0	1		+		
1	1	1	1				+

COM is internally ties to a GND

Differential MUX Mode

Address				Channel #			
SGL/ $\overline{\text{DIF}}$	$\overline{\text{ODD}}$/ $\overline{\text{SIGN}}$	SELECT 1	SELECT 0	0	1	2	3
0	0	0	1	+	−		
0	0	1	1			+	−
0	1	0	1	−	+		
0	1	1	1			−	+

Functional Description (Continued)

Since the input configuration is under software control, it can be modified, as required, at each conversion. A channel can be treated as a single-ended, ground referenced input for one conversion; then it can be reconfigured as part of a differential channel for another conversion. *Figure 1* illustrates the input flexibility which can be achieved.

The analog input voltages for each channel can range from 50 mV below ground to 50mV above V_{CC}(typically 5V) without degrading conversion accuracy.

2.0 THE DIGITAL INTERFACE

A most important characteristic of these converters is their serial data link with the controlling processor. Using a serial communication format offers two very significant system improvements; it allows more function to be included in the converter package with no increase in package size and it can eliminate the transmission of low level analog signals by locating the converter right at the analog sensor; transmitting highly noise immune digital data back to the host processor.

To understand the operation of these converters it is best to refer to the Timing Diagram and Functional Block Diagram and to follow a complete conversion sequence.

1. A conversion is initiated by first pulling the $\overline{CS}$ (chip select) line low. This line must be held low for the entire conversion. The converter is now waiting for a start bit and its MUX assignment word.

2. A clock is then generated by the processor (if not provided continuously) and output to the A/D clock input.

3. On each rising edge of the clock the status of the data in (DI) line is clocked into the MUX address shift register. The start bit is the first logic "1" that appears on this line (all leading zeros are ignored). Following the start bit the converter expects the next 4 bits to be the MUX assignment word.

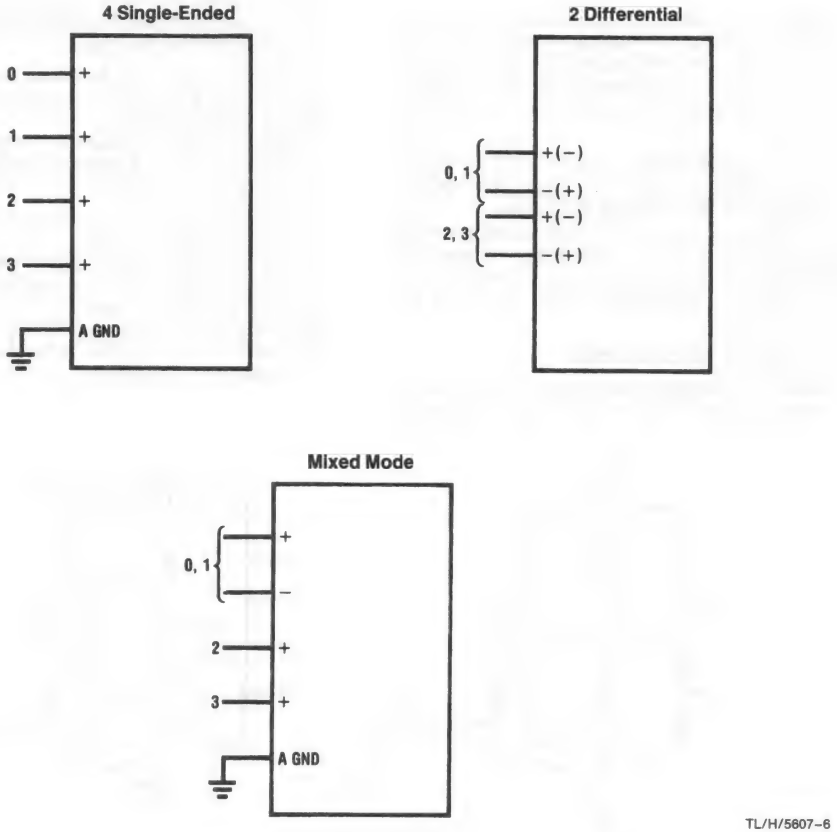

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FIGURE 1. Analog Input Multiplexer Options for the ADC0833

Functional Description (Continued)

4. When the start bit has been shifted into the start location of the MUX register, the input channel has been assigned and a conversion is about to begin. An interval of ½ clock period (where nothing happens) is automatically inserted to allow the selected MUX channel to settle. The SAR status line goes high at this time to signal that a conversion is now in progress and the DI line is disabled (it no longer accepts data).

5. The data out (DO) line now comes out of TRI-STATE and provides a leading zero for this one clock period of MUX settling time.

6. When the conversion begins, the output of the SAR comparator, which indicates whether the analog input is greater than (high) or less than (low) each successive voltage from the internal resistor ladder, appears at the DO line on each falling edge of the clock. This data is the result of the conversion being shifted out (with the MSB coming first) and can be read by the processor immediately.

7. After 8 clock periods the conversion is completed. The SAR status line returns low to indicate this ½ clock cycle later.

8. If the programmer prefers, the data can be read in an LSB first format. All 8 bits of the result are stored in an output shift register. The conversion result, LSB first, is automatically shifted out the DO line, after the MSB first data stream. The DO line then goes low and stays low until $\overline{CS}$ is returned high.

9. All internal registers are cleared when the $\overline{CS}$ line is high. If another conversion is desired, $\overline{CS}$ must make a high to low transition followed by address information.

The DI and DO lines can be tied together and controlled through a bidirectional processor I/O bit with one wire. This is possible because the DI input is only "looked-at" during the MUX addressing interval while the DO line is still in a high impedance state.

3.0 REFERENCE CONSIDERATIONS

The ADC0833 is intended primarily for use in circuits requiring absolute accuracy. In this type of system, the analog inputs vary between very specific voltage limits and the reference voltage for the A/D converter must remain stable with time and temperature. For ratiometric applications, an ADC0834 is a pin-for-pin compatible alternative since it has a V_{REF} input (note the ADC0834 needs one less bit of mux addressing information).

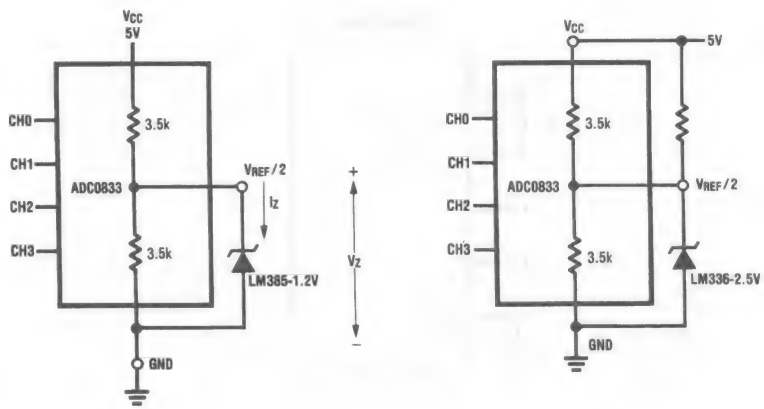
The voltage applied to the $V_{REF}/2$ pin defines the voltage span of the analog input [the difference between $V_{IN}(+)$ and $V_{IN}(-)$] over which the 256 possible output codes apply. A full-scale conversion (an all 1s output code) will result when the voltage difference between a selected "+" input and "−" input is approximately *twice* the voltage at the $V_{REF}/2$ pin. This internal gain of 2 from the applied reference to the full-scale input voltage allows biasing a low voltage reference diode from the $5V_{DC}$ converter supply. To accommodate a 5V input span, only a 2.5V reference is required. The LM385 and LM336 reference diodes are good low current devices to use with these converters. The output code changes in accordance with the following equation:

$$\text{Output Code} = 256 \left(\frac{V_{IN}(+) - V_{IN}(-)}{2(V_{REF}/2)} \right)$$

where the output code is the decimal equivalent of the 8-bit binary output (ranging from 0 to 255) and the term $V_{REF}/2$ is the voltage from pin 9 to ground.

The $V_{REF}/2$ pin is the center point of a two resistor divider (each resistor is 3.5 kΩ) connected from V_{CC} to ground. Total ladder input resistance is the sum of these two equal resistors. As shown in *Figure 2,* a reference diode with a voltage less than $V_{CC}/2$ can be connected without requiring an external biasing resistor if its current requirements meet the indicated level.

The minimum value of $V_{REF}/2$ can be quite small (see Typical Performance Characteristics) to allow direct conversions of transducer outputs providing less than a 5V output span. Particular care must be taken with regard to noise pickup, circuit layout and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter (1 LSB equals $V_{REF}/256$).



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$V_{FULL\text{-}SCALE} \cong 2.4V$ $\qquad$ $V_{FULL\text{-}SCALE} \cong 5.0V$

Note: No external biasing resistor needed if $V_Z < \frac{V_{CC}}{2}$ and I_Z min $< \frac{V_{CC}/2 - V_Z}{1.75\ k\Omega}$

FIGURE 2. Reference Biasing Examples

Functional Description (Continued)

4.0 THE ANALOG INPUTS

The most important feature of these converters is that they can be located right at the analog signal source and through just a few wires can communicate with a controlling processor with a highly noise immune serial bit stream. This in itself greatly minimizes circuitry to maintain analog signal accuracy which otherwise is most susceptible to noise pickup. However, a few words are in order with regard to the analog inputs should the inputs be noisy to begin with or possibly riding on a large common-mode voltage.

The differential input of these converters actually reduces the effects of common-mode input noise, a signal common to both selected "+" and "−" inputs for a conversion (60 Hz is most typical). The time interval between sampling the "+" input and then the "−" input is ½ of a clock period. The change in the common-mode voltage during this short time interval can cause conversion errors. For a sinusoidal common-mode signal this error is:

$$V_{error}(max) = V_{PEAK}(2\pi f_{CM})\left(\frac{0.5}{f_{CLK}}\right)$$

where f_{CM} is the frequency of the common-mode signal,

V_{PEAK} is its peak voltage value

and f_{CLK} is the A/D clock frequency.

For a 60 Hz common-mode signal to generate a ¼ LSB error (≈5 mV) with the converter running at 250 kHz, its peak value would have to be 6.63V which would be larger than allowed as it exceeds the maximum analog input limits.

Due to the sampling nature of the analog inputs short spikes of current enter the "+" input and exit the "−" input at the clock edges during the actual conversion. These currents decay rapidly and do not cause errors as the internal comparator is strobed at the end of a clock period. Bypass capacitors at the inputs will average these currents and cause an effective DC current to flow through the output resistance of the analog signal source. Bypass capacitors should not be used if the source resistance is greater than 1 kΩ.

This source resistance limitation is important with regard to the DC leakage currents of input multiplexer as well. The worst-case leakage current of ± 1 μA over temperature will create a 1 mV inut error with a 1 kΩ source resistance. An op amp RC active low pass filter can provide both impedance buffering and noise filtering should a high impedance signal source be required.

5.0 OPTIONAL ADJUSTMENTS

5.1 Zero Error

The zero of the A/D does not require adjustment. If the minimum analog input voltage value, $V_{IN(MIN)}$, is not ground a zero offset can be done. The converter can be made to output 0000 0000 digital code for this minimum input voltage by biasing any V_{IN} (−) input at this $V_{IN(MIN)}$ value. This utilizes the differential mode operation of the A/D.

The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the $V_{IN}(-)$ input and applying a small magnitude positive voltage to the $V_{IN}(+)$ input. Zero error is the difference between the actual DC input voltage which is necessary to just cause an output digital code transition from 0000 0000 to 0000 0001 and the ideal ½ LSB value (½ LSB = 9.8 mV for $V_{REF}/2 = 2.500\ V_{DC}$).

5.2 Full-Scale

The full-scale adjustment can be made by applying a differential input voltage which is 1 ½ LSB down from the desired analog full-scale voltage range and then adjusting the magnitude of the V_{REF} input or V_{CC} for a digital output code which is just changing from 1111 1110 to 1111 1111.

5.3 Adjusting for an Arbitrary Analog Input Voltage Range

If the analog zero voltage of the A/D is shifted away from ground (for example, to accommodate an analog input signal which does not go to ground), this new zero reference should be properly adjusted first. A $V_{IN}(+)$ voltage which equals this desired zero reference plus ½ LSB (where the LSB is calculated for the desired analog span, using 1 LSB = analog span/256) is applied to selected "+" input and the zero reference voltage at the corresponding "−" input should then be adjusted to just obtain the 00_{HEX} to 01_{HEX} code transition.

The full-scale adjustment should be made [with the proper $V_{IN}(-)$ voltage applied] by forcing a voltage to the $V_{IN}(+)$ input which is given by:

$$V_{IN}(+)\ fs\ adj = V_{MAX} - 1.5\left[\frac{(V_{MAX} - V_{MIN})}{256}\right]$$

where:

V_{MAX} = the high end of the analog input range

and

V_{MIN} = the low end (the offset zero) of the analog range.

(Both are ground referenced.)

The $V_{REF}/2$ voltage is then adjusted to provide a code change from FE_{HEX} to FF_{HEX}. This completes the adjustment procedure.

6.0 POWER SUPPLY

A unique feature of the ADC0833 is the inclusion of a 7V zener diode connected from the V^+ terminal to ground which also connects to the V_{CC} terminal (which is the actual converter supply) through a silicon diode, as shown in *Figure 3.*

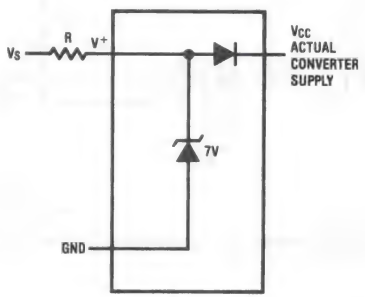

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FIGURE 3. An On-Chip Shunt Regulator Diode

Functional Description (Continued)

This zener is intended for use as a shunt voltage regulator to eliminate the need for any additional regulating components. This is most desirable if the converter is to be remotely located from the system power source. *Figures 4* and *5* illustrate two useful applications of this on-board zener when an external transistor can be afforded.

An important use of the interconnecting diode between V^+ and V_{CC} is shown in *Figures 6* and *7*. Here, this diode is used as a rectifier to allow the V_{CC} supply for the converter to be derived from the clock. The low current requirements of the A/D (~3 mA) and the relatively high clock frequencies used (typically in the range of 10k-400 kHz) allows using the small value filter capacitor shown to keep the ripple on the V_{CC} line to well under ¼ of an LSB. The shunt zener regulator can also be used in this mode. This requires a clock voltage swing which is in excess of V_Z. A current limit for the zener is needed, either built into the clock generator or a resistor can be used from the CLK pin to the V^+ pin.

Applications

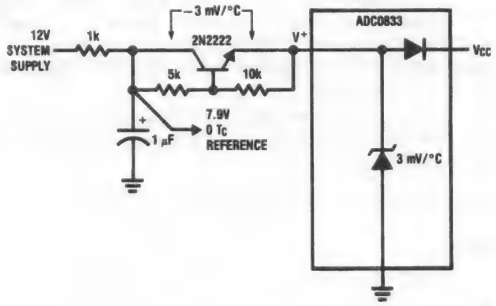

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FIGURE 4. Operating with a Temperature Compensated Reference

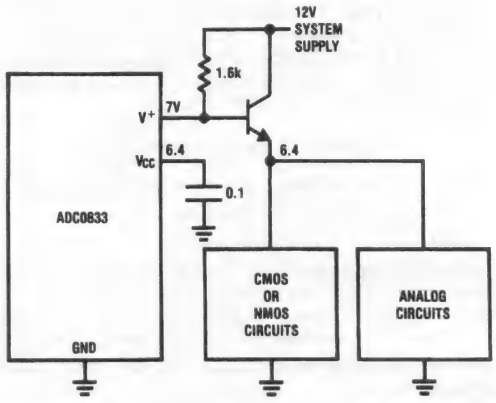

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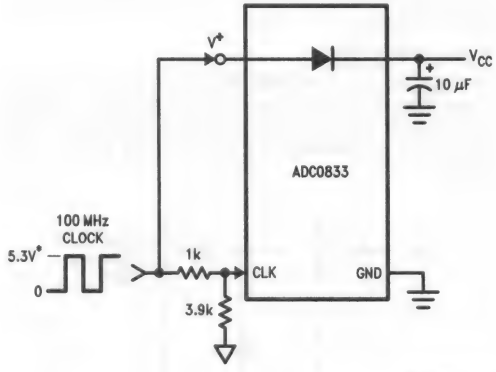
FIGURE 5. Using the A/D as the System Supply Regulator



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*Note 4.5V ≤ V_{CC} ≤ 6.3V

FIGURE 6. Generally V_{CC} from the Converter Clock

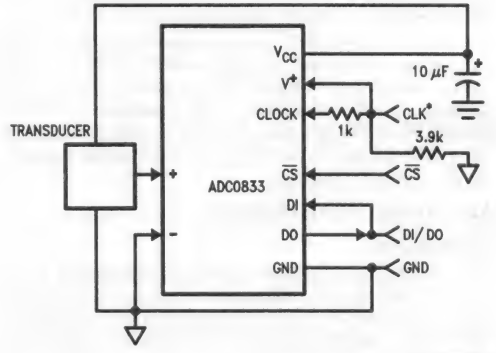

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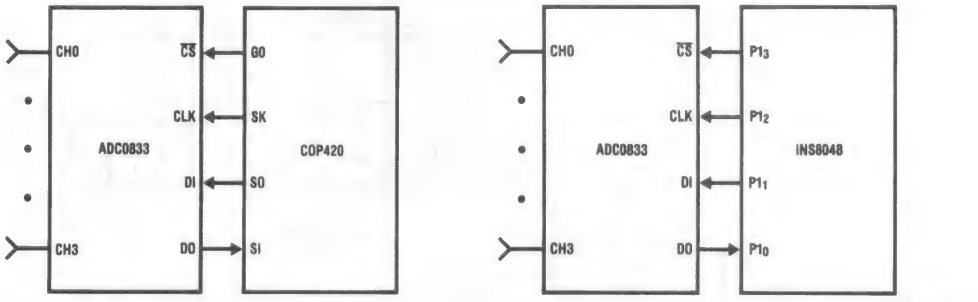
FIGURE 7. Remote Sensing—Clock and Power on 1 Wire

Applications (Continued)

Digital Link and Sample Controlling Software for the Serially Oriented COP420 and the Bit Programmable I/O INS8048



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COP CODING EXAMPLE

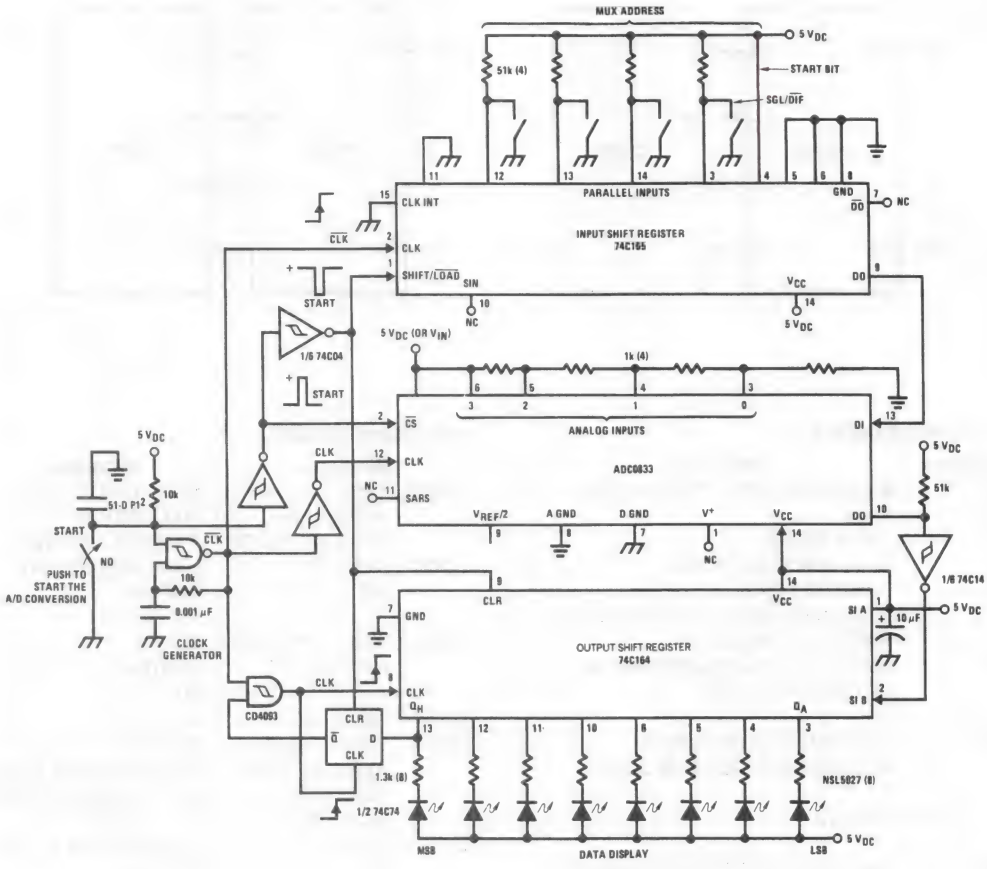
Mnemonic	Instruction
LEI	ENABLES SIO's INPUT AND OUTPUT
SC	C = 1
OGI	G0 = 0 ($\overline{CS}$ = 0)
CLR A	CLEARS ACCUMULATOR
AISC 1	LOADS ACCUMULATOR WITH 1
XAS	EXCHANGES SIO WITH ACCUMULATOR AND STARTS SK CLOCK
LDD	LOADS MUX ADDRESS FROM RAM INTO ACCUMULATOR
NOP	—
XAS	LOADS MUX ADDRESS FROM ACCUMULATOR TO SIO REGISTER
↑ 8 INSTRUCTIONS ↓	
XAS	READS HIGH ORDER NIBBLE (4 BITS) INTO ACCUMULATOR
XIS	PUTS HIGH ORDER NIBBLE INTO RAM
CLR A	CLEARS ACCUMULATOR
RC	C = 0
XAS	READS LOW ORDER NIBBLE INTO ACCUMULATOR AND STOPS SK
XIS	PUTS LOW ORDER NIBBLE INTO RAM
OGI	G0 = 1 ($\overline{CS}$ = 1)
LEI	DISABLES SIO's INPUT AND OUTPUT

8048 CODING EXAMPLE

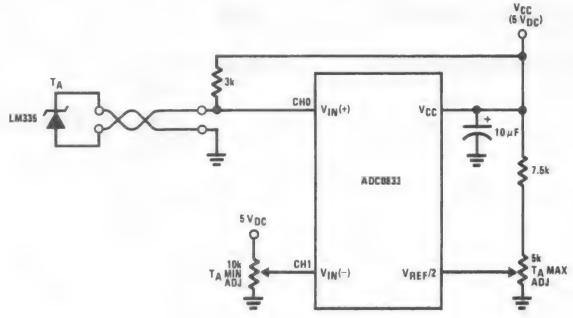
	Mnemonic		Instruction
START:	ANL	P1, #0F7H	;SELECT A/D ($\overline{CS}$ = 0)
	MOV	B, #5	;BIT COUNTER ← 5
	MOV	A, #ADDR	;A ← MUX ADDRESS
LOOP 1:	RRC	A	;CY ← ADDRESS BIT
	JC	ONE	;TEST BIT
			;BIT = 0
ZERO:	ANL	P1, #0FEH	;DI ← 0
	JMP	CONT	;CONTINUE
			;BIT = 1
ONE:	ORL	P1, #1	;DI ← 1
CONT:	CALL	PULSE	;PULSE SK 0 → 1 → 0
	DJNZ	B, LOOP 1	;CONTINUE UNTIL DONE
	CALL	PULSE	;EXTRA CLOCK FOR SYNC
	MOV	B, #8	;BIT COUNTER ← 8
LOOP 2:	CALL	PULSE	;PULSE SK 0 → 1 → 0
	IN	A, P1	;CY ← DO
	RRC	A	
	RRC	A	
	MOV	A, C	;A ← RESULT
	RLC	A	;A(0) ← BIT AND SHIFT
	MOV	C, A	;C ← RESULT
	DJNZ	B, LOOP 2	;CONTINUE UNTIL DONE
RETR			
			;PULSE SUBROUTINE
PULSE:	ORL	P1, #04	;SK ← 1
	NOP		;DELAY
	ANL	P1, #0FBH	;SK ← 0
	RET		

Applications (Continued)

A "Stand-Alone" Hook-Up for ADC0833 Evaluation



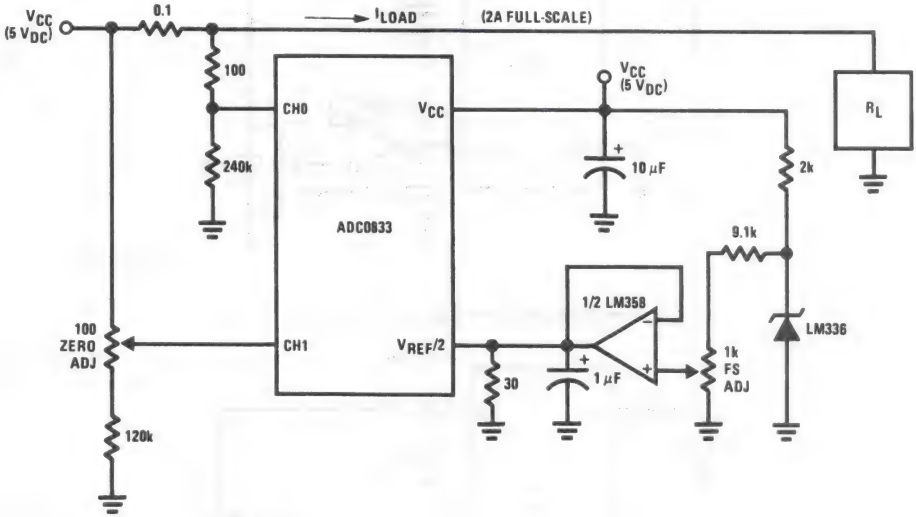
Low Cost Remote Temperature Sensor



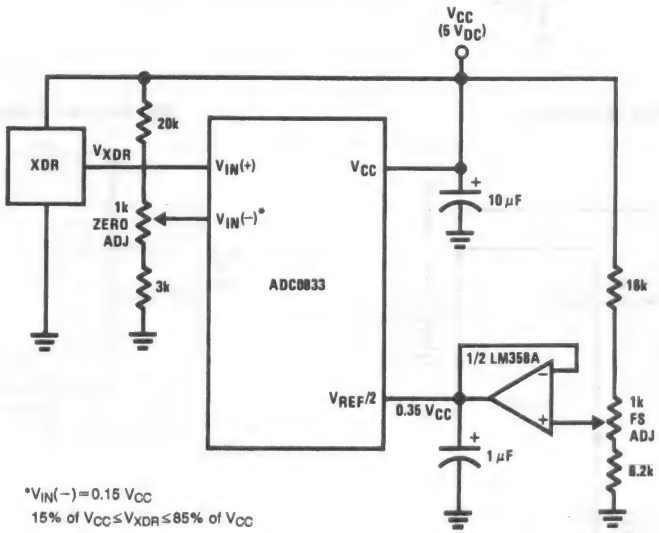
TL/H/5607–11

Applications (Continued)

Digitizing a Current Flow



Operating with Automotive Ratiometric Transducers

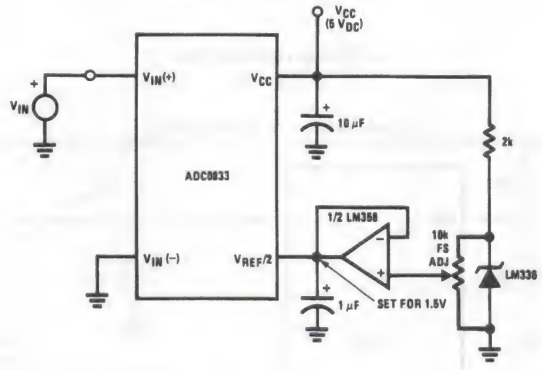


*$V_{IN}(-) = 0.15\ V_{CC}$
15% of $V_{CC} \leq V_{XDR} \leq 85\%$ of V_{CC}

TL/H/5607–12

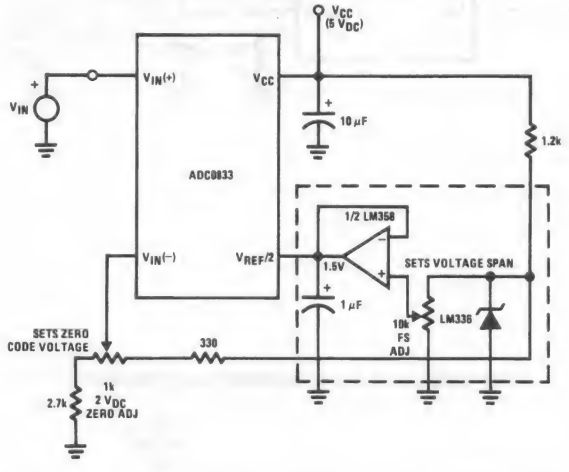
Applications (Continued)

Span Adjust: $0V \leq V_{IN} \leq 3V$



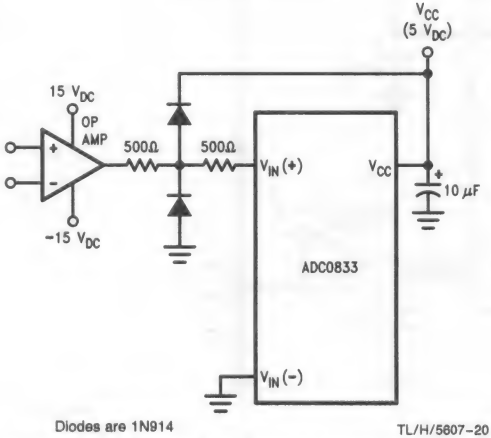
TL/H/5607–18

Zero-Shift and Span Adjust: $2V \leq V_{IN} \leq 5V$



TL/H/5607–19

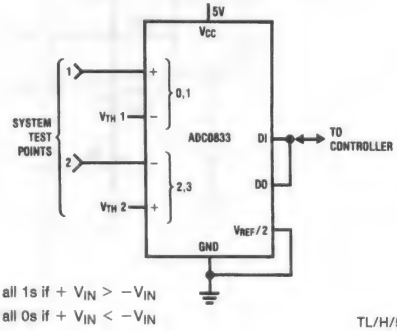
Protecting the Input



Diodes are 1N914

TL/H/5607–20

High Accuracy Comparators



DO = all 1s if + V_{IN} > − V_{IN}

DO = all 0s if + V_{IN} < − V_{IN}

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For additional application ideas, refer to the data sheet for the ADC0831 family of serial data converters.

Ordering Information

Part Number	Temperature Range	Total Unadjusted Error
ADC0833BCJ	−40°C to +85°C	±1/2 LSB
ADC0833BCN	0°C to +70°C	
ADC0833BJ	−55°C to +125°C	
ADC0833CCJ	−40°C to +85°C	±1 LSB
ADC0833CCN	0°C to +70°C	
ADC0833CJ	−55°C to +125°C	

National Semiconductor

ADC0841 8-Bit μP Compatible A/D Converter

General Description

The ADC0841 is a CMOS 8-bit successive approximation A/D converter. Differential inputs provide low frequency input common mode rejection and allow offsetting the analog range of the converter. In addition, the reference input can be adjusted enabling the conversion of reduced analog ranges with 8-bit resolution.

The A/D is designed to operate with the control bus of a variety of microprocessors. TRI-STATE® output latches that directly drive the data bus permit the A/D to be configured as a memory location or I/O device to the microprocessor with no interface logic necessary.

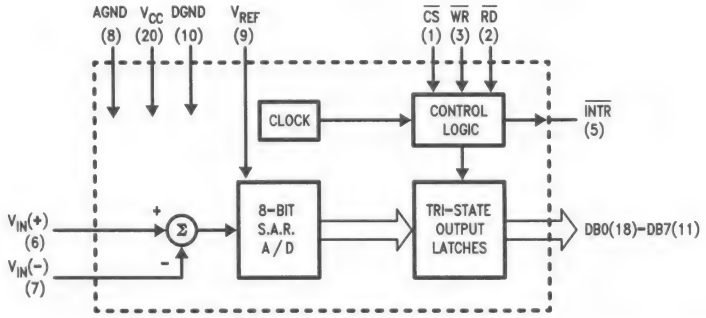
Features

- Easy interface to all microprocessors
- Operates ratiometrically or with 5 V_{DC} voltage reference
- No zero or full-scale adjust required
- Internal clock
- 0V to 5V input range with single 5V power supply
- 0.3" standard width 20-pin package
- 20 Pin Molded Chip Carrier Package

Key Specifications

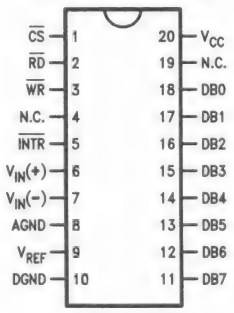
■ Resolution	8 Bits
■ Total Unadjusted Error	±½ LSB and ± 1 LSB
■ Single Supply	5 V_{DC}
■ Low Power	15 mW
■ Conversion Time	40 μs

Block and Connection Diagrams



TL/H/8557–1

Dual-In-Line Package

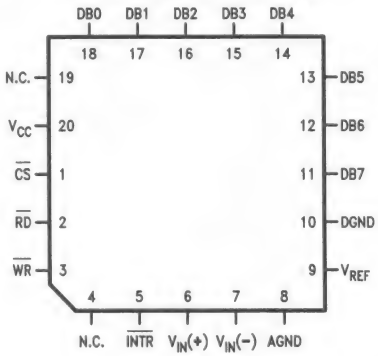


TL/H/8557–2

Top View

(N.C.-No Connection)

Molded Chip Carrier Package



TL/H/8557–3

Top View

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	6.5V
Voltage	
Logic Control Inputs	$-0.3V$ to $V_{CC}+0.3V$
At Other Inputs and Outputs	$-0.3V$ to $V_{CC}+0.3V$
Input Current Per Pin (Note 3)	±5 mA
Input Current Per Package (Note 3)	±20 mA
Storage Temperature	−65°C to +150°C
Package Dissipation at $T_A=25°C$	875 mW
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C
ESD Susceptibility (Note 10)	800V

Operating Conditions (Notes 1 and 2)

Supply Voltage (V_{CC})	4.5 V_{DC} to 6.0 V_{DC}
Temperature Range	$T_{MIN}\le T_A\le T_{MAX}$
ADC0841BCN, ADC0841CCN	$0°C\le T_A\le 70°C$
ADC0841BCJ, ADC0841CCJ, ADC0841BCV, ADC0841CCV	$-40°C\le T_A\le 85°C$
ADC0841BJ, ADC0841CJ	$-55°C\le T_A\le 125°C$

Electrical Characteristics

The following specifications apply for $V_{CC}=5$ V_{DC} unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A=T_j=25°C$.

Parameter	Conditions	ADC0841BJ, ADC0841BCJ ADC0841CJ, ADC0841CCJ			ADC0841BCN, ADC0841CCN ADC0841BCV, ADC0841CCV			Units
		Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	
CONVERTER AND MULTIPLEXER CHARACTERISTICS								
Maximum Total Unadjusted Error	$V_{REF}=5.00$ V_{DC} (Note 4)							
ADC0841BCN, BCV						±½	**±½**	LSB
ADC0841BJ, BCJ			**±½**					LSB
ADC0841CCN, CCV						±1	**±1**	LSB
ADC0841CJ, CCJ			**±1**					LSB
Minimum Reference Input Resistance		2.4	**1.1**		2.4	1.2	**1.1**	kΩ
Maximum Reference Input Resistance		2.4	**5.9**		2.4	5.4	**5.9**	kΩ
Maximum Common-Mode Input Voltage	(Note 5)		**$V_{CC}+0.05$**			$V_{CC}+0.05$	**$V_{CC}+0.05$**	V
Minimum Common-Mode Input Voltage	(Note 5)		**GND−0.05**			GND−0.05	**GND−0.05**	V
DC Common-Mode Error	Differential Mode	±1/16	**±¼**		±1/16	±¼	**±¼**	LSB
Power Supply Sensitivity	$V_{CC}=5V\pm5\%$	±1/16	**±⅛**		±1/16	±⅛	**±⅛**	LSB

Electrical Characteristics

The following specifications apply for $V_{CC} = 5\ V_{DC}$ unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_j = 25°C$. (Continued)

Symbol	Parameter	Conditions	ADC0841BJ, ADC0841BCJ ADC0841CJ, ADC0841CCJ			ADC0841BCN, ADC0841CCN ADC0841BCV, ADC0841CCV			Units
			Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	
DIGITAL AND DC CHARACTERISTICS									
$V_{IN(1)}$	Logical "1" Input Voltage (Min)	$V_{CC} = 5.25V$		**2.0**			2.0	**2.0**	V
$V_{IN(0)}$	Logical "0" Input Voltage (Max)	$V_{CC} = 4.75V$		**0.8**			0.8	**0.8**	V
$I_{IN(1)}$	Logical "1" Input Current (Max)	$V_{IN} = 5.0V$	0.005	**1**		0.005		**1**	μA
$I_{IN(0)}$	Logical "0" Input Current (Max)	$V_{IN} = 0V$	−0.005	**−1**		−0.005		**−1**	μA
$V_{OUT(1)}$	Logical "1" Output Voltage (Min)	$V_{CC} = 4.75V$ $I_{OUT} = -360\ \mu A$ $I_{OUT} = -10\ \mu A$		 **2.4** **4.5**			 2.8 4.6	 **2.4** **4.5**	 V V
$V_{OUT(0)}$	Logical "0" Output Voltage (Max)	$V_{CC} = 4.75V$ $I_{OUT} = 1.6\ mA$		**0.4**			0.34	**0.4**	V
I_{OUT}	TRI-STATE Output Current (Max)	$V_{OUT} = 0V$ $V_{OUT} = 5V$	−0.01 0.01	**−3** **3**		−0.01 0.01	−0.3 0.3	**−3** **3**	μA μA
I_{SOURCE}	Output Source Current (Min)	$V_{OUT} = 0V$	−14	**−6.5**		−14	−7.5	**−6.5**	mA
I_{SINK}	Output Sink Current (Min)	$V_{OUT} = V_{CC}$	16	**8.0**		16	9.0	**8.0**	mA
I_{CC}	Supply Current (Max)	$\overline{CS} = 1$, V_{REF} Open	1	**2.5**		1	2.3	**2.5**	mA

AC Characteristics

The following specifications apply for $V_{CC} = 5V_{DC}$, $t_r = t_f = 10$ ns unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Symbol	Parameter	Conditions	Typ (Note 6)	Tested Limit (Note 7)	Design Limit (Note 8)	Units
t_C	Maximum Conversion Time (See Graph)		30	40	**60**	μs
$t_{W(\overline{WR})}$	Minimum $\overline{WR}$ Pulse Width	(Note 9)	50	150		ns
t_{ACC}	Maximum Access Time (Delay from Falling Edge of $\overline{RD}$ to Output Data Valid)	$C_L = 100$ pF (Note 9)	145	225		ns
t_{1H}, t_{0H}	TRI-STATE Control (Maximum Delay from Rising Edge of $\overline{RD}$ to Hi-Z State)	$C_L = 10$ pF, $R_L = 10$k, $t_r = 20$ ns (Note 9)	125		200	ns
t_{WI}, t_{RI}	Maximum Delay from Falling Edge of $\overline{WR}$ or $\overline{RD}$ to Reset of $\overline{INTR}$	(Note 9)	200	400		ns
C_{IN}	Capacitance of Logic Inputs		5			pF
C_{OUT}	Capacitance of Logic Outputs		5			pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to the ground pins.

Note 3: During over-voltage conditions ($V_{IN} < 0V$ and $V_{IN} > V_{CC}$) the maximum input current at any one pin is ±5 mA. If the current is limited to ±5 mA at all the pins no more than four pins can be in this condition in order to meet the Input Current Per Package (±20 mA) specification.

Note 4: Total undajusted error includes offset, full-scale, and linearity.

Note 5: For $V_{IN}(-) \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input, which will forward-conduct for analog input voltages one diode drop below ground or one diode drop greater than V_{CC} supply. Be careful during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct, especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

Note 6: Typicals are at 25°C and represent most likely parametric norm.

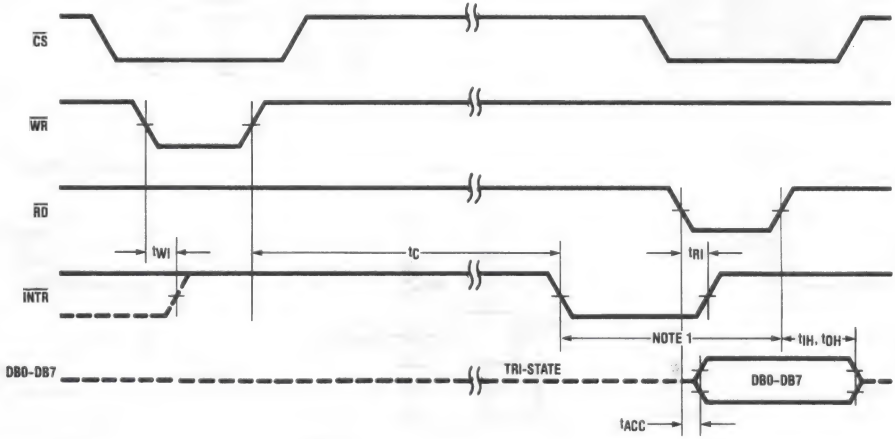
Note 7: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 8: Design limits are guaranteed but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Note 9: The temperature coefficient is 0.3%/°C.

Note 10: Human body model, 100 pF discharged through 1.5 kΩ resistor.

Timing Diagram

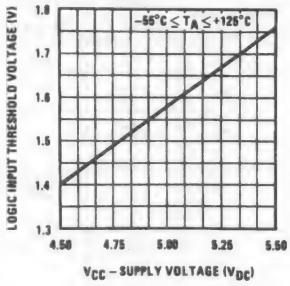


TL/H/8557–9

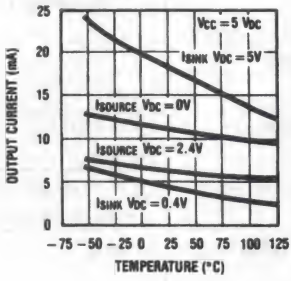
Note 1: Read strobe must occur at least 600 ns after the assertion of interrupt to guarantee reset of $\overline{INTR}$.

Typical Performance Characteristics

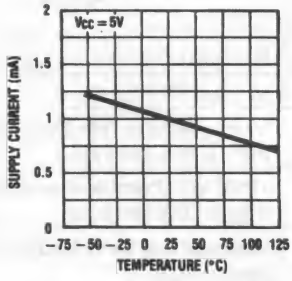
Logic Input Threshold Voltage vs Supply Voltage



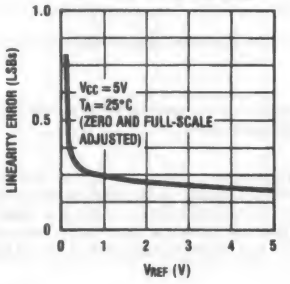
Output Current vs Temperature



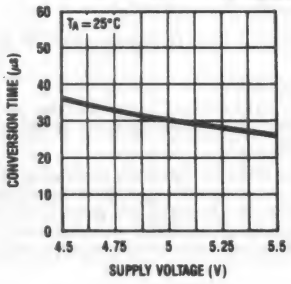
Power Supply Current vs Temperature



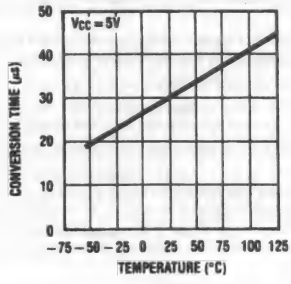
Linearity Error vs V_{REF}



Conversion Time vs V_{SUPPLY}

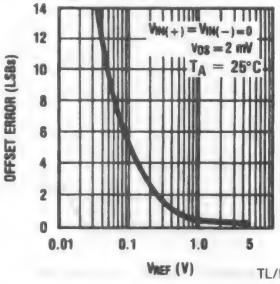


Conversion Time vs Temperature



TL/H/8557–4

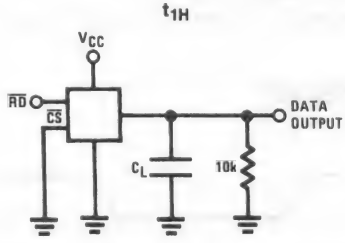
Unadjusted Offset Error vs V_{REF} Voltage



TL/H/8557–22

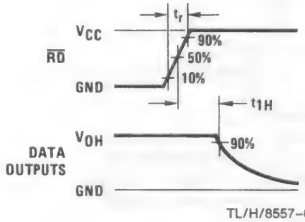
TRI-STATE Test Circuits and Waveforms

t_{1H}



TL/H/8557–5

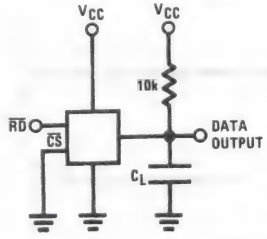
t_{1H}, C_L = 10 pF



TL/H/8557–6

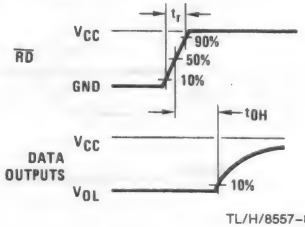
t_r = 20 ns

t_{0H}



TL/H/8557–7

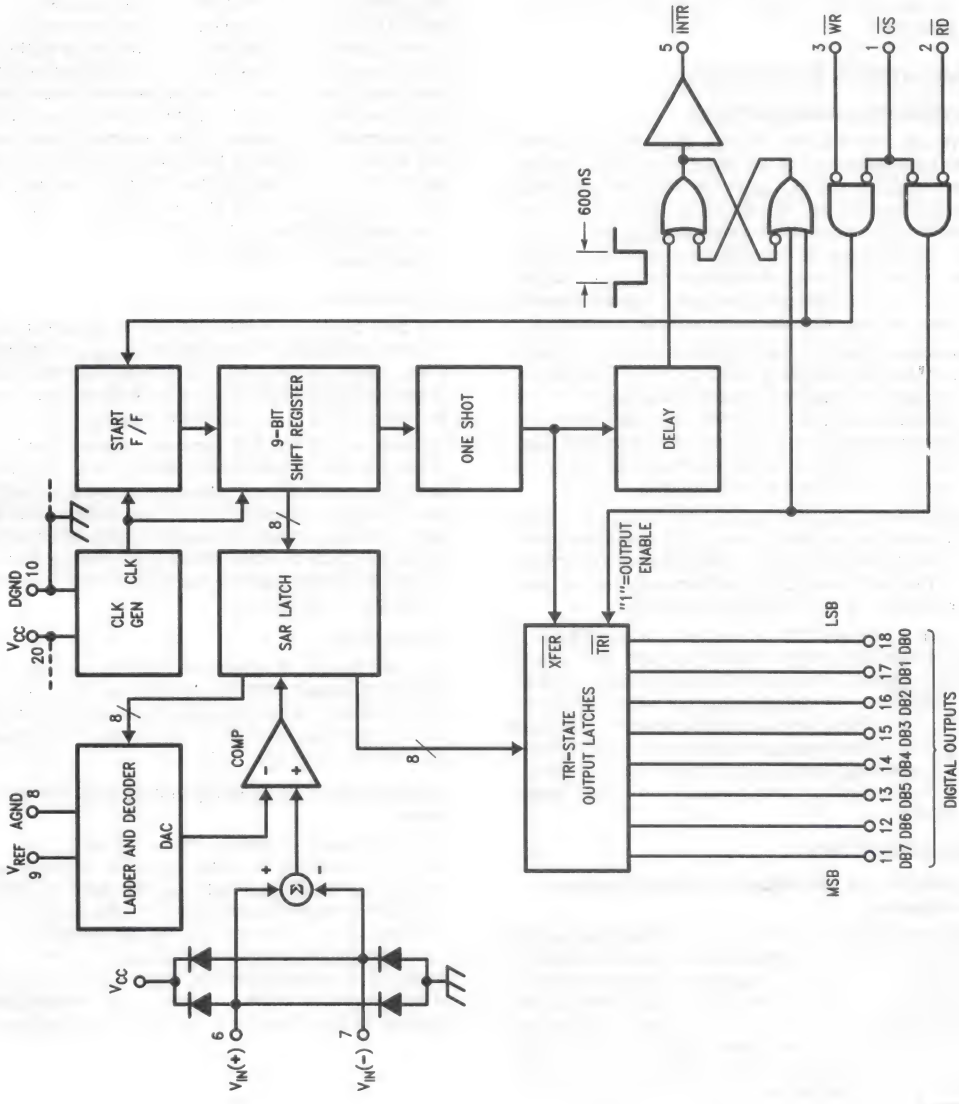
t_{0H}, C_L = 10 pF



TL/H/8557–8

t_r = 20 ns

Functional Block Diagram



TL/H/8557–10

Functional Description

A conversion is initiated via the $\overline{CS}$ and $\overline{WR}$ lines. If the data from a previous conversion is not read, the $\overline{INTR}$ line will be low. The falling edge of $\overline{WR}$ will reset the $\overline{INTR}$ line high and ready the A/D for a conversion cycle. The rising edge of $\overline{WR}$ starts a conversion. After the conversion cycle ($t_C \leq 60$ μsec), which is set by the internal clock frequency, the digital data is transferred to the output latch and the $\overline{INTR}$ is asserted low. Taking $\overline{CS}$ and $\overline{RD}$ low resets $\overline{INTR}$ output high and transfers the conversion result on the output data lines (DB0–DB7).

Applications Information

1.0 REFERENCE CONSIDERATIONS

The voltage applied to the reference input of this converter defines the voltage span of the analog input (the difference between $V_{IN(MAX)}$ and $V_{IN(MIN)}$) over which the 256 possible output codes apply. The device can be used in either ratiometric applications or in systems requiring absolute accuracy. The reference pin must be connected to a voltage source capable of driving the minimum reference input resistance of 1.1 kΩ. This pin is the top of a resistor divider string used for the successive approximation conversion.

In a ratiometric system (*Figure 1a*), the analog input voltage is proportional to the voltage used for the A/D reference. This voltage is typically the system power supply, so the V_{REF} pin can be tied to V_{CC}. This technique relaxes the stability requirements of the system reference as the analog input and A/D reference move together maintaining the same output code for a given input condition.

For absolute accuracy (*Figure 1b*), where the analog input varies between very specific voltage limits, the reference pin can be biased with a time and temperature stable voltage source. The LM385 and LM336 reference diodes are good low current devices to use with this converter.

The maximum value of the reference is limited to the V_{CC} supply voltage. The minimum value, however, can be quite small (see Typical Performance Characteristics) to allow direct conversions of transducer outputs providing less than a 5V output span. Particular care must be taken with regard to noise pickup, circuit layout and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter (1 LSB equals $V_{REF}/256$).

2.0 THE ANALOG INPUTS

2.1 Analog Differential Voltage Inputs and Common-Mode Rejection

The differential inputs of this converter actually reduce the effects of common-mode input noise, a signal common to both selected "+" and "−" inputs for a conversion (60 Hz is most typical). The time interval between sampling the "+" input and then the "−" input is ½ of a clock period. The change in the common-mode voltage during this short time interval can cause conversion errors. For a sinusoidal common-mode signal this error is:

$$V_{ERROR(MAX)} = V_{peak}\,(2\pi\, f_{CM}) \times 0.5 \times \left(\frac{t_C}{8}\right)$$

where f_{CM} is the frequency of the common-mode signal, Vpeak is its peak voltage value and t_C is the conversion time.

For a 60 Hz common-mode signal to generate a ¼ LSB error (≈5 mV) with the converter running at 40 μS, its peak value would have to be 5.43V. This large a common-mode signal is much greater than that generally found in a well designed data acquisition system.

2.2 Input Current

Due to the sampling nature of the analog inputs, short duration spikes of current enter the "+" input and exit the "−" input at the clock edges during the actual conversion. These currents decay rapidly and do not cause errors as the internal comparator is strobed at the end of a clock period. Bypass capacitors at the inputs will average these currents and cause an effective DC current to flow through the output resistance of the analog signal source. Bypass capacitors should not be used if the source resistance is greater than 1 kΩ. An op amp RC active low pass filter can provide both impedance buffering and noise filtering should a high impedance signal source be required.

3.0 OPTIONAL ADJUSTMENTS

3.1 Zero Error

The zero of the A/D does not require adjustment. If the minimum analog input voltage value, $V_{IN(MIN)}$, is not ground, a zero offset can be done. The converter can be made to output 0000 0000 digital code for this minimum input voltage by biasing the V_{IN} (−) input at this $V_{IN(MIN)}$ value.

The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the V^- input and applying a small magnitude positive voltage to the V^+ input. Zero error is the difference between actual DC input voltage which is necessary to just cause an output digital code transition from 0000 0000 to 0000 0001 and the ideal ½ LSB value (½ LSB = 9.8 mV for $V_{REF} = 5.000\ V_{DC}$).

3.2 Full-Scale

The full-scale adjustment can be made by applying a differential input voltage which is 1 ½ LSB down from the desired analog full-scale voltage range and then adjusting the magnitude of the V_{REF} input for a digital output code changing from 1111 1110 to 1111 1111.

3.3 Adjusting for an Arbitrary Analog Input Voltage Range

If the analog zero voltage of the A/D is shifted away from ground (for example, to accommodate an analog input signal which does not go to ground), this new zero reference should be properly adjusted first. A voltage which equals this desired zero reference plus ½ LSB (where the LSB is calculated for the desired analog span, 1 LSB = analog span/256) is applied to the "+" input ($V_{IN}(+)$) and the zero reference voltage at the "−" input ($V_{IN}(-)$)should then be adjusted to just obtain the 00_{HEX} to 01_{HEX} code transition.

Applications Information (Continued)

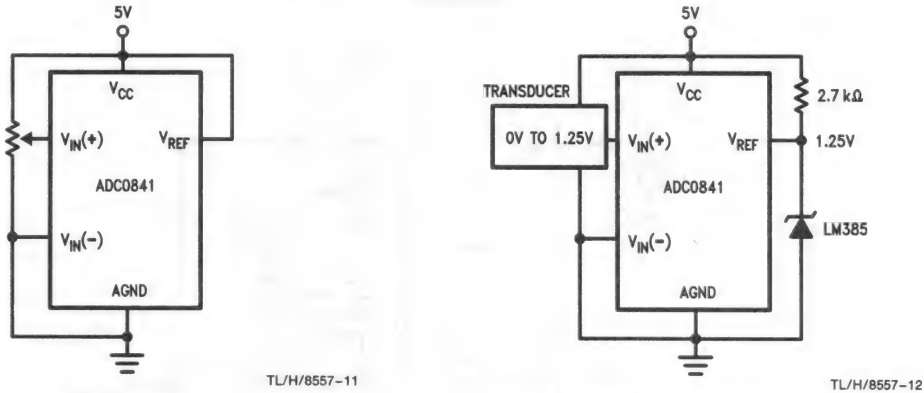


FIGURE 1. Referencing Examples

The full-scale adjustment should be made [with the proper $V_{IN}(-)$ voltage applied] by forcing a voltage to the $V_{IN}(+)$ input which is given by:

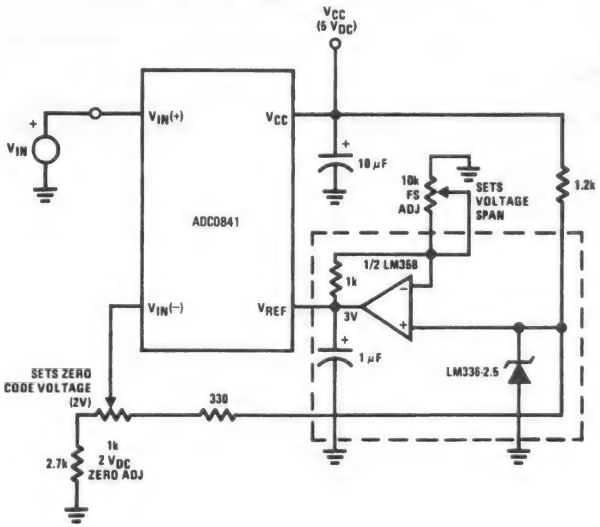
$$V_{IN}(+)\,\text{fs adj} = V_{MAX} - 1.5\left[\frac{(V_{MAX} - V_{MIN})}{256}\right]$$

where V_{MAX} = the high end of the analog input range and V_{MIN} = the low end (the offset zero) of the analog range. (Both are ground referenced.)

The V_{REF} (or V_{CC}) voltage is then adjusted to provide a code change from FE_{HEX} to FF_{HEX}. This completes the adjustment procedure.

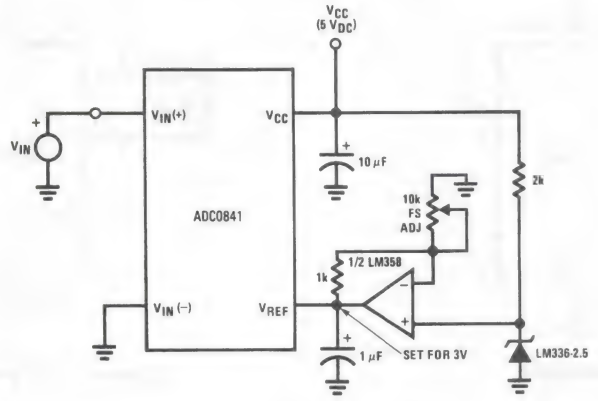
For an example see the Zero-Shift and Span Adjust circuit below.

Zero-Shift and Span Adjust ($2V \leq V_{IN} \leq 5V$)



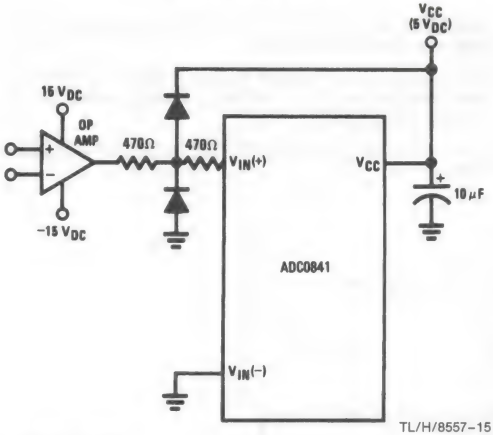
Applications Information (Continued)

Span Adjust $0V \leq V_{IN} \leq 3V$



TL/H/8557–14

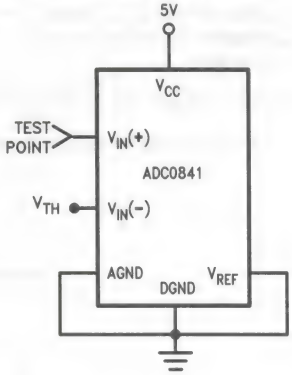
Protecting the Input



TL/H/8557–15

Diodes are 1N914

High Accuracy Comparator

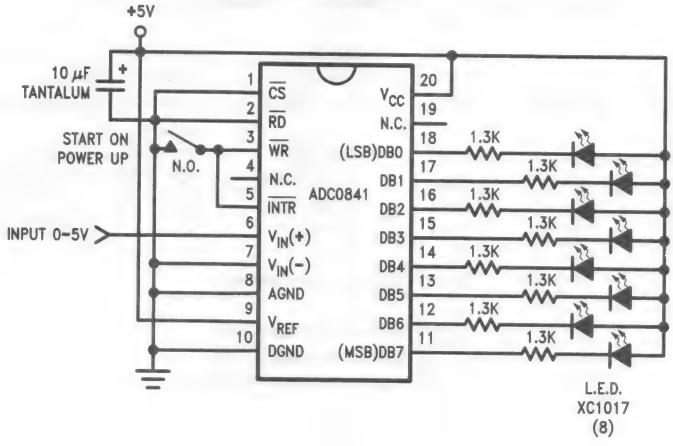


TL/H/8557–16

DO = all 1s if $V_{IN}(+) > V_{IN}(-)$
DO = all 0s if $V_{IN}(+) < V_{IN}(-)$

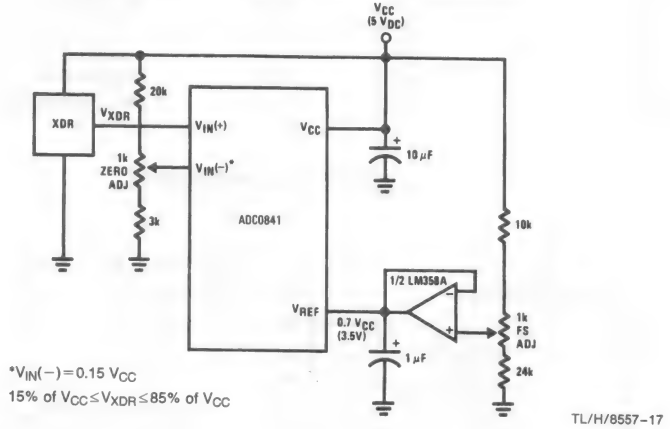
Applications Information (Continued)

Continuous Conversion



TL/H/8557–19

Operating with Automotive Ratiometric Transducers



*$V_{IN}(-) = 0.15\ V_{CC}$

15% of $V_{CC} \leq V_{XDR} \leq 85\%$ of V_{CC}

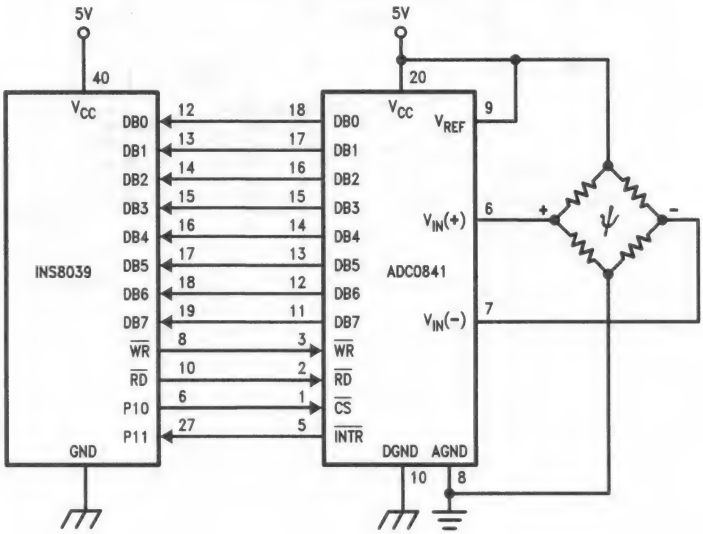
TL/H/8557–17

Applications Information (Continued)

SAMPLE PROGRAM FOR ADC0841—INS8039 INTERFACE
CONVERTING TWO RATIOMETRIC, DIFFERENTIAL SIGNALS

```
                                ORG     0H
0000    04 10                   JMP     BEGIN     ;START PROGRAM AT ADDR 10
                                ORG     10H       ;MAIN PROGRAM
0010    B9 FF           BEGIN:  MOV     R1,#0FFH  ;LOAD R1 WITH A UNUSED ADDR
                                                  ;LOCATION
0012    B8 20                   MOV     R0,#20H   ;A/D DATA ADDRESS
0014    89 FF                   ORL     P1,#0FFH  ;SET PORT 1 OUTPUTS HIGH
0016    23 00                   MOV     A,00H     ;LOAD THE ACC WITH 00
0018    14 50                   CALL    CONV      ;CALL THE CONVERSION SUBROUTINE

                                ;CONTINUE MAIN PROGRAM

                                ;CONVERSION SUBROUTINE
                                ;ENTRY:ACC—A/D MUX DATA
                                ;EXIT: ACC—CONVERTED DATA

                                ORG     50H
0050    99 FE           CONV:   ANL     P1,#0FEH  ;CHIP SELECT THE A/D
0052    91                      MOVX    @R1,A     ;START CONVERSION
0053    09              LOOP:   IN      A,P1      ;INPUT INTR STATE
0054    32 53                   JB1     LOOP      ;IF INTR = 1 GOTO LOOP
0056    81                      MOVX    A,@R1     ;IF INTR = 0 INPUT A/D DATA
0057    89 01                   ORL     P1,&01H   ;CLEAR THE A/D CHIP SELECT
0059    A0                      MOV     @R0,A     ;STORE THE A/D DATA
005A    83                      RET               ;RETURN TO MAIN PROGRAM
```

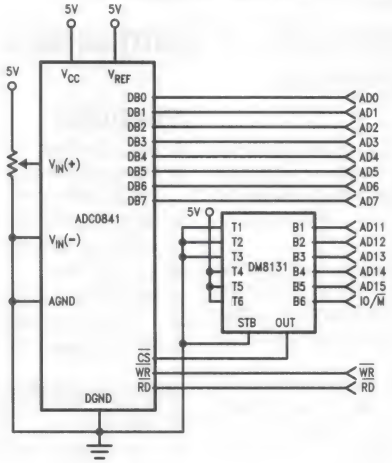
ADC0841—INS8039 Interface



TL/H/8557–20

Applications Information (Continued)

I/O Interface to NSC800™



TL/H/8557–21

SAMPLE PROGRAM FOR ADC0841—NSC800 INTERFACE

```
0010                NCONV   EQU   16          ;TWICE THE NUMBER OF REQUIRED
                                                ;CONVERSIONS
000F                DEL     EQU   15          ;DELAY 60 μsec CONVERSION
001F                CS      EQU   1FH         ;THE BOARD ADDRESS
3C00                ADDTA   EQU   003CH       ;START OF RAM FOR A/D
                                                ;DATA
0000'  00           DTA:    DB    08H         ; DATA
0001'  0E 1F        START:  LD    C,CS
0003'  06 16                LD    B,NCONV
0005'  21 0000'             LD    HL,DTA
0008'  11 003C              LD    DE,ADDTA
000B'  ED A3        STCONV: OUTI                ;START A CONVERSION
000D'  EB                   EX    DE,HL       ;HL = RAM ADDRESS FOR THE
                                                ;A/D DATA
000E'  3E 0F                LD    A,DEL
0010'  3D           WAIT:   DEC   A           ;WAIT 60 μsec FOR THE
0011'  C2 0013'             JP    NZ,WAIT     ;CONVERSION TO FINISH
0014'  ED A2                INI                 ;STORE THE A/D'S DATA
                                                ;THE REQUIRED CONVERSIONS COMPLETED?
0016'  EB                   EX    DE,HL
0017'  C2 000E'             JP    NZ,STCONV   ;IF NOT GOTO STCONV

                            END
```

Note: A conversion is started, then a 60 μs wait for the A/D to complete a conversion and the data is stored at address ADDTA for the first conversion, ADDTA + 1 for the second conversion, etc. for a total of 8 conversions.

Ordering Information

Temperature Range	Total Unadjusted Error		Package Outline
	±½ LSB	±1 LSB	
0°C to +70°C	ADC0841BCN	ADC0841CCN	N20A Molded Dip
−40°C to +85°C	ADC0841BCJ	ADC0841CCJ	J20A Cerdip
	ADC0841BCV	ADC0841CCV	V20A Molded Chip Carrier
−55°C to +125°C	ADC0841BJ	ADC0841CJ	J20A Cerdip

3

National Semiconductor

ADC0844/ADC0848 8-Bit μP Compatible A/D Converters with Multiplexer Options

General Description

The ADC0844 and ADC0848 are CMOS 8-bit successive approximation A/D converters with versatile analog input multiplexers. The 4-channel or 8-channel multiplexers can be software configured for single-ended, differential or pseudo-differential modes of operation.

The differential mode provides low frequency input common mode rejection and allows offsetting the analog range of the converter. In addition, the A/D's reference can be adjusted enabling the conversion of reduced analog ranges with 8-bit resolution.

The A/Ds are designed to operate from the control bus of a wide variety of microprocessors. TRI-STATE® output latches that directly drive the data bus permit the A/Ds to be configured as memory locations or I/O devices to the microprocessor with no interface logic necessary.

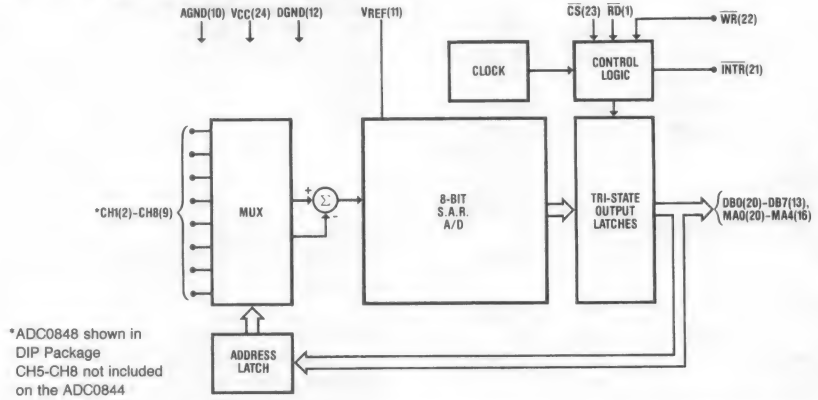
Features

- Easy interface to all microprocessors
- Operates ratiometrically or with 5 V_{DC} voltage reference
- No zero or full-scale adjust required
- 4-channel or 8-channel multiplexer with address logic
- Internal clock
- 0V to 5V input range with single 5V power supply
- 0.3" standard width 20-pin or 24-pin DIP
- 28 Pin Molded Chip Carrier Package

Key Specifications

■ Resolution	8 Bits
■ Total Unadjusted Error	±½ LSB and ± 1 LSB
■ Single Supply	5 V_{DC}
■ Low Power	15 mW
■ Conversion Time	40 μs

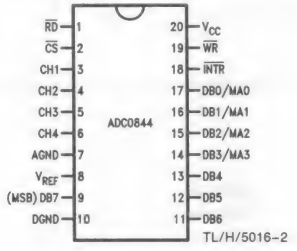
Block and Connection Diagrams



*ADC0848 shown in DIP Package
CH5-CH8 not included on the ADC0844

TL/H/5016-1

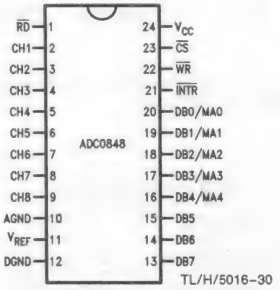
Dual-In-Line Package



TL/H/5016-2

Top View

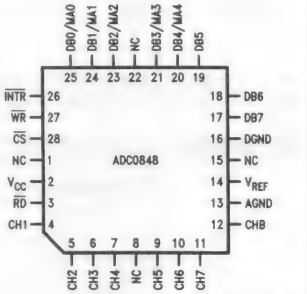
Dual-In-Line Package



TL/H/5016-30

Top View

Molded Chip Carrier Package



TL/H/5016-29

Top View

See Ordering Information

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	6.5V
Voltage	
Logic Control Inputs	−0.3V to +15V
At Other Inputs and Outputs	−0.3V to V_{CC}+0.3V
Input Current at Any Pin (Note 3)	5 mA
Package Input Current (Note 3)	20 mA
Storage Temperature	−65°C to +150°C
Package Dissipation at T_A=25°C	875 mW
ESD Susceptibility (Note 4)	800V
Lead Temperature (Soldering, 10 seconds)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C

Operating Conditions (Notes 1 & 2)

Supply Voltage (V_{CC})	4.5 V_{DC} to 6.0 V_{DC}
Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0844BCN, ADC0844CCN, ADC0848BCN, ADC0848CCN	0°C≤T_A≤70°C
ADC0844BCJ, ADC0844CCJ, ADC0848BCJ, ADC0848CCJ ADC0848BCV, ADC0848CCV	−40°C≤T_A≤85°C
ADC0844BJ, ADC0844CJ, ADC0848BJ, ADC0848CJ	−55°C≤T_A≤125°C

Electrical Characteristics

The following specifications apply for V_{CC} = 5 V_{DC} unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits T_A = T_j = 25°C.

Parameter	Conditions	ADC0844BJ, ADC0844BCJ ADC0844CJ, ADC0844CCJ ADC0848BJ, ADC0848BCJ ADC0848CJ, ADC0848CCJ			ADC0844BCN, ADC0844CCN ADC0848BCN, ADC0848CCN ADC0848BCV, ADC0848CCV			Limit Units
		Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	
CONVERTER AND MULTIPLEXER CHARACTERISTICS								
Maximum Total Unadjusted Error	V_{REF}=5.00 V_{DC} (Note 8)							
ADC0844BCN, ADC0848BCN, BCV						±½	**±½**	LSB
ADC0844BJ, BCJ, ADC0848BJ, BCJ			**±½**					LSB
ADC0844CCN, ADC0848CCN, CCV						±1	**±1**	LSB
ADC0844CJ, CCJ, ADC0848CJ, CCJ			**±1**					LSB
Minimum Reference Input Resistance		2.4	**1.1**		2.4	1.2	**1.1**	kΩ
Maximum Reference Input Resistance		2.4	**5.9**		2.4	5.4	**5.9**	kΩ
Maximum Common-Mode Input Voltage	(Note 9)		**V_{CC}+0.05**			V_{CC}+0.05	**V_{CC}+0.05**	V
Minimum Common-Mode Input Voltage	(Note 9)		**GND−0.05**			GND−0.05	**GND−0.05**	V
DC Common-Mode Error	Differential Mode	±1/16	**±¼**		±1/16	±¼	**±¼**	LSB
Power Supply Sensitivity	V_{CC}=5V±5%	±1/16	**±⅛**		±1/16	±⅛	**±⅛**	LSB
Off Channel Leakage Current	(Note 10) On Channel=5V, Off Channel=0V		**−1**			−0.1	**−1**	µA
	On Channel=0V, Off Channel=5V		**1**			0.1	**1**	µA
DIGITAL AND DC CHARACTERISTICS								
$V_{IN(1)}$, Logical "1" Input Voltage (Min)	V_{CC}=5.25V		**2.0**			2.0	**2.0**	V
$V_{IN(0)}$, Logical "0" Input Voltage (Max)	V_{CC}=4.75V		**0.8**			0.8	**0.8**	V
$I_{IN(1)}$, Logical "1" Input Current (Max)	V_{IN}=5.0V	0.005	**1**		0.005		**1**	µA

Electrical Characteristics

The following specifications apply for V_{CC} = 5 V_{DC} unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_j$ = 25°C. (Continued)

Parameter	Conditions	ADC0844BJ, ADC0844BCJ ADC0844CJ, ADC0844CCJ ADC0848BJ, ADC0848BCJ ADC0848CJ, ADC0848CCJ			ADC0844BCN, ADC0844CCN ADC0848BCN, ADC0848CCN ADC0848BCV, ADC0848CCV			Limit Units
		Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	
DIGITAL AND DC CHARACTERISTICS (Continued)								
$I_{IN(0)}$, Logical "0" Input Current (Max)	V_{IN} = 0V	−0.005	**−1**		−0.005		**−1**	μA
$V_{OUT(1)}$, Logical "1" Output Voltage (Min)	V_{CC} = 4.75V I_{OUT} = −360 μA I_{OUT} = −10 μA		**2.4** **4.5**			2.8 4.6	**2.4** **4.5**	V V
$V_{OUT(0)}$, Logical "0" Output Voltage (Max)	V_{CC} = 4.75V I_{OUT} = 1.6 mA		**0.4**			0.34	**0.4**	V
I_{OUT}, TRI-STATE Output Current (Max)	V_{OUT} = 0V V_{OUT} = 5V	−0.01 0.01	**−3** **3**		−0.01 0..01	−0.3 0.3	**−3** **3**	μA μA
I_{SOURCE}, Output Source Current (Min)	V_{OUT} = 0V	−14	**−6.5**		−14	−7.5	**−6.5**	mA
I_{SNK}, Output Sink Current (Min)	V_{OUT} = V_{CC}	16	**8.0**		16	9.0	**8.0**	mA
I_{CC}, Supply Current (Max)	$\overline{CS}$ = 1, V_{REF} Open	1	**2.5**		1	2.3	**2.5**	mA

AC Electrical Characteristics

The following specifications apply for V_{CC} = 5V_{DC}, $t_r = t_f$ = 10 ns unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_j$ = 25°C.

Parameter	Conditions	Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	Units
t_C, Maximum Conversion Time (See Graph)		30	40	**60**	μs
$t_{W(\overline{WR})}$, Minimum $\overline{WR}$ Pulse Width	(Note 11)	50	150		ns
t_{ACC}, Maximum Access Time (Delay from Falling Edge of $\overline{RD}$ to Output Data Valid)	C_L = 100 pF (Note 11)	145		225	ns
t_{1H}, t_{0H}, TRI-STATE Control (Maximum Delay from Rising Edge of $\overline{RD}$ to Hi-Z State)	C_L = 10 pF, R_L = 10k (Note 11)	125		200	ns
t_{WI}, t_{RI}, Maximum Delay from Falling Edge of $\overline{WR}$ or $\overline{RD}$ to Reset of $\overline{INTR}$	(Note 11)	200	400		ns
t_{DS}, Minimum Data Set-Up Time	(Note 11)	50	100		ns
t_{DH}, Minimum Data Hold Time	(Note 11)	0	50		ns
C_{IN}, Capacitance of Logic Inputs		5			pF
C_{OUT}, Capacitance of Logic Outputs		5			pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to the ground pins.

Note 3: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of the current at that pin should be limited to 5 mA or less. The 20 mA package input current limits the number of pins that can exceed the power supply boundaries with a 5 mA current limit to four.

Note 4: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 5: Typicals are at 25°C and represent most likely parametric norm.

Note 6: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 7: Design limits are guaranteed by not 100% tested. These limits are not used to calculate outgoing quality levels.

Note 8: Total unadjusted error includes offset, full-scale, linearity, and multiplexer error.

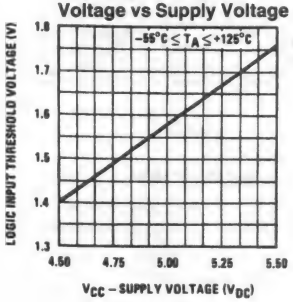
Note 9: For $V_{IN}(-) \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input, which will forward-conduct for analog input voltages one diode drop below ground or one diode drop greater than V_{CC} supply. Be careful during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct, especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

Note 10: Off channel leakage current is measured after the channel selection.

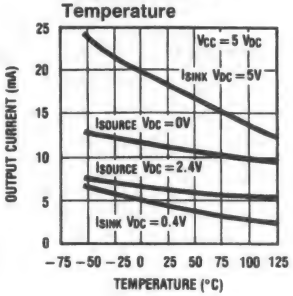
Note 11: The temperature coefficient is 0.3%/°C.

Typical Performance Characteristics

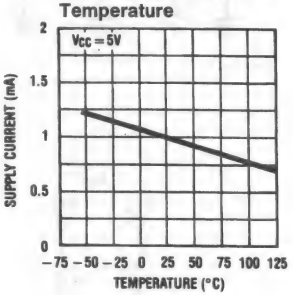
Logic Input Threshold Voltage vs Supply Voltage



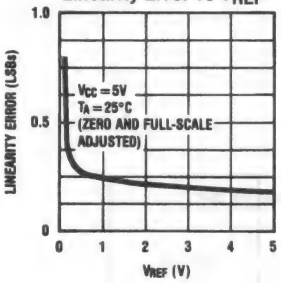
Output Current vs Temperature



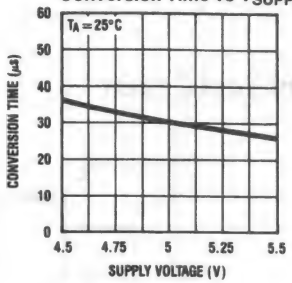
Power Supply Current vs Temperature



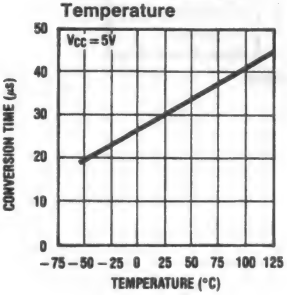
Linearity Error vs V_{REF}



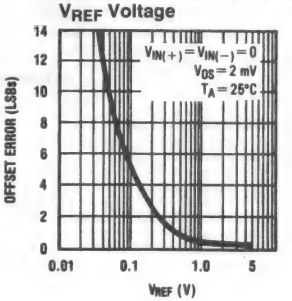
Conversion Time vs V_{SUPPLY}



Conversion Time vs Temperature



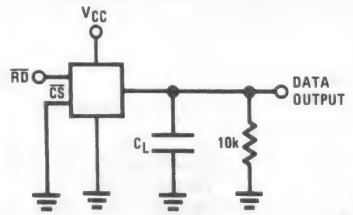
Unadjusted Offset Error vs V_{REF} Voltage



TL/H/5016–3

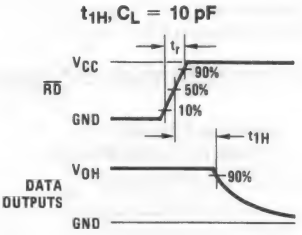
TRI-STATE Test Circuits and Waveforms

t_{1H}



TL/H/5016-4

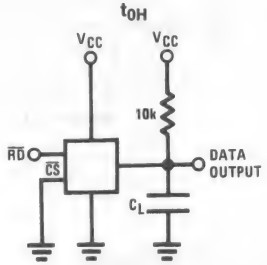
t_{1H}, C_L = 10 pF



TL/H/5016-5

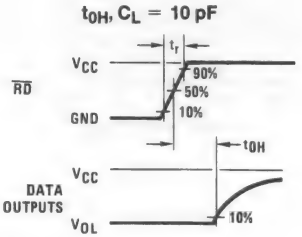
t_r = 20 ns

t_{0H}



TL/H/5016-6

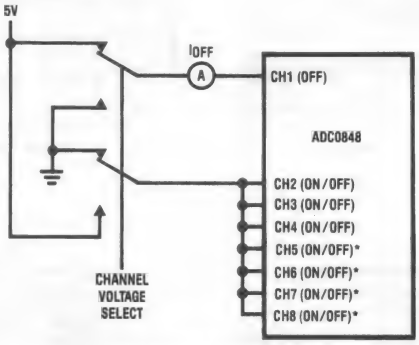
t_{0H}, C_L = 10 pF



TL/H/5016-7

t_r = 20 ns

Leakage Current Test Circuit

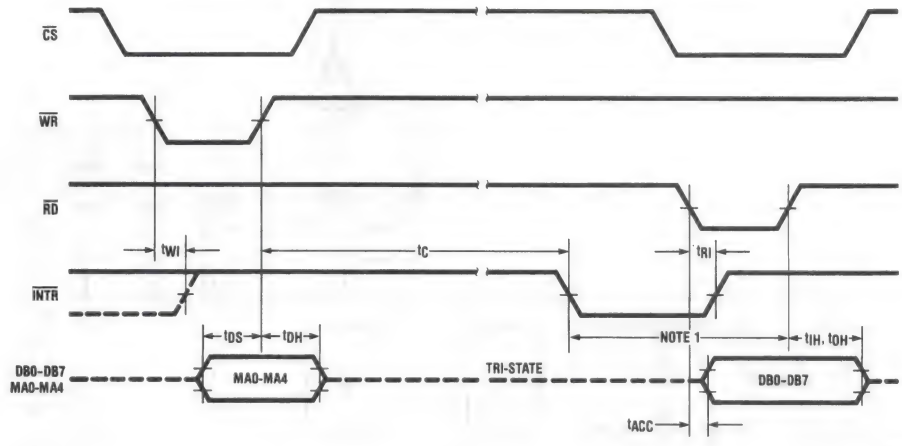


*NOT INCLUDED ON ADC0844

TL/H/5016-8

Timing Diagrams

Programming New Channel Configuration and Starting a Conversion

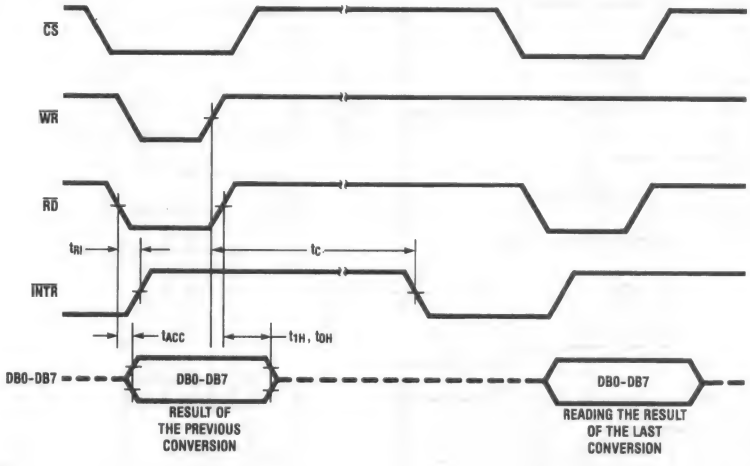


TL/H/5016–9

Note 1: Read strobe must occur at least 600 ns after the assertion of interrupt to guarantee reset of $\overline{INTR}$.

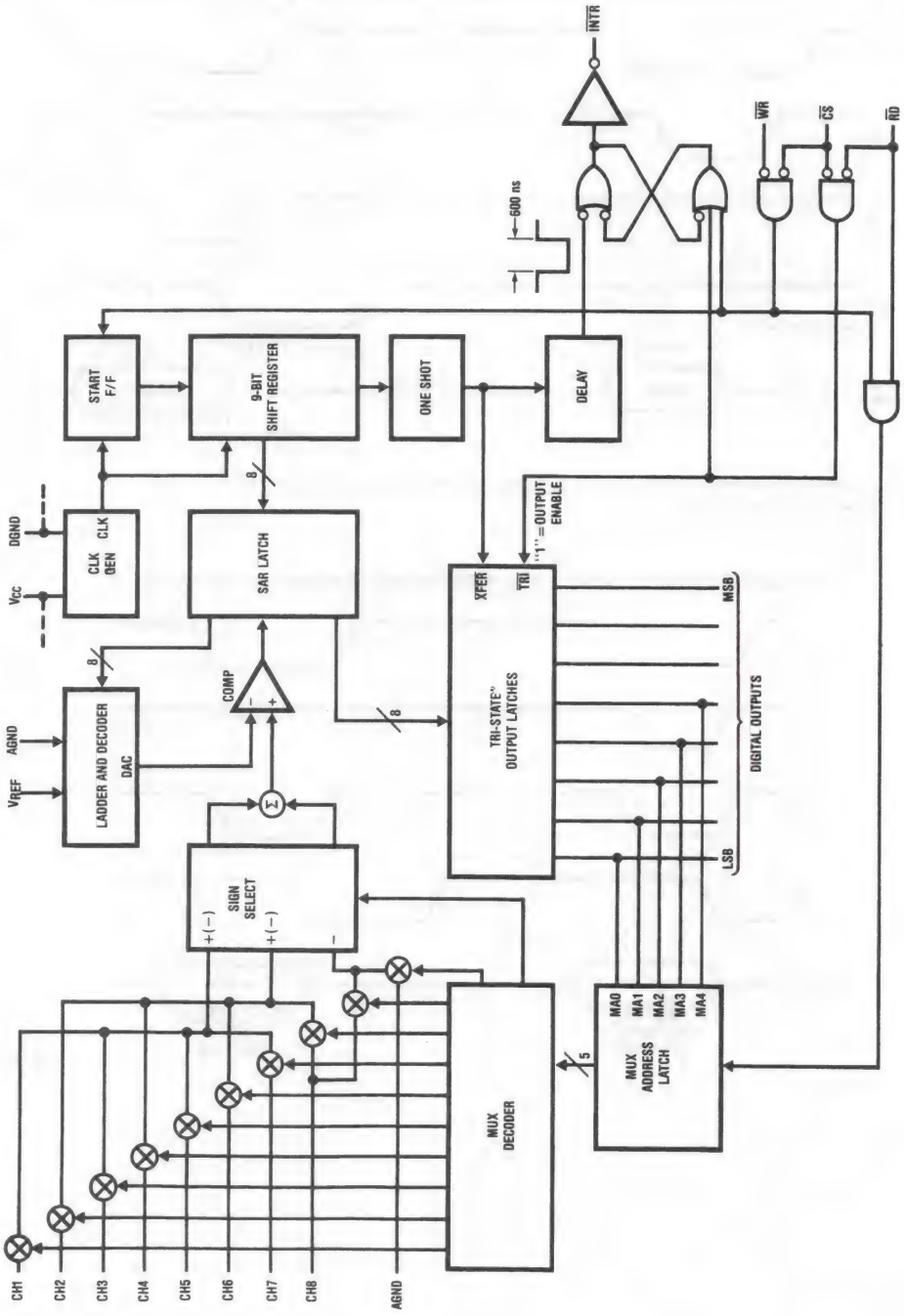
Note 2: MA stands for MUX address.

Using the Previously Selected Channel Configuration and Starting a Conversion



TL/H/5016–10

ADC0848 Functional Block Diagram



TL/H/5016–11

Functional Description

The ADC0844 and ADC0848 contain a 4-channel and 8-channel analog input multiplexer (MUX) respectively. Each MUX can be configured into one of three modes of operation differential, pseudo-differential, and single ended. These modes are discussed in the Applications Information Section. The specific mode is selected by loading the MUX address latch with the proper address (see Table I and Table II). Inputs to the MUX address latch (MA0-MA4) are common with data bus lines (DB0-DB4) and are enabled when the $\overline{RD}$ line is high. A conversion is initiated via the $\overline{CS}$ and $\overline{WR}$ lines. If the data from a previous conversion is not read, the $\overline{INTR}$ line will be low. The falling edge of $\overline{WR}$ will reset the $\overline{INTR}$ line high and ready the A/D for a conversion cycle. The rising edge of $\overline{WR}$, with $\overline{RD}$ high, strobes the data on the MA0/DB0-MA4/DB4 inputs into the MUX address latch to select a new input configuration and start a conversion. If the $\overline{RD}$ line is held low during the entire low period of $\overline{WR}$ the previous MUX configuration is retained, and the data of the previous conversion is the output on lines DB0-DB7. After the conversion cycle ($t_C \leq 40\ \mu s$), which is set by the internal clock frequency, the digital data is transferred to the output latch and the $\overline{INTR}$ is asserted low. Taking $\overline{CS}$ and $\overline{RD}$ low resets $\overline{INTR}$ output high and outputs the conversion result on the data lines (DB0-DB7).

Applications Information

1.0 MULTIPLEXER CONFIGURATION

The design of these converters utilizes a sampled-data comparator structure which allows a differential analog input to be converted by a successive approximation routine.

The actual voltage converted is always the difference between an assigned "+" input terminal and a "−" input terminal. The polarity of each input terminal of the pair being converted indicates which line the converter expects to be the most positive. If the assigned "+" input is less than the "−" input the converter responds with an all zeros output code.

A unique input multiplexing scheme has been utilized to provide multiple analog channels. The input channels can be software configured into three modes: differential, single-

TABLE I. ADC0844 MUX ADDRESSING

MUX Address				$\overline{CS}$	$\overline{WR}$	$\overline{RD}$	Channel #					MUX Mode
MA3	MA2	MA1	MA0				CH1	CH2	CH3	CH4	AGND	
X	L	L	L	L	↓	H	+	−				Differential
X	L	L	H	L		H	−	+				
X	L	H	L	L		H			+	−		
X	L	H	H	L		H			−	+		
L	H	L	L	L	↓	H	+				−	Single-Ended
L	H	L	H	L		H		+			−	
L	H	H	L	L		H			+		−	
L	H	H	H	L		H				+	−	
H	H	L	L	L	↓	H	+			−		Pseudo-Differential
H	H	L	H	L		H		+		−		
H	H	H	L	L		H			+	−		
X	X	X	X	L	↓	L	Previous Channel Configuration					

X = don't care

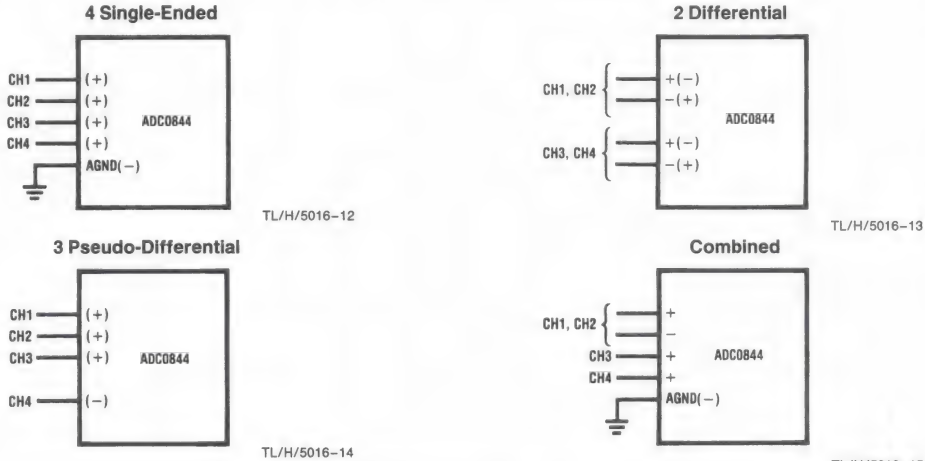


FIGURE 1. Analog Input Multiplexer Options

Applications Information (Continued)

ended, or pseudo-differential. *Figure 1* shows the three modes using the 4-channel MUX ADC0844. The eight inputs of the ADC0848 can also be configured in any of the three modes. In the differential mode, the ADC0844 channel inputs are grouped in pairs, CH1 with CH2 and CH3 with CH4. The polarity assignment of each channel in the pair is interchangeable. The single-ended mode has CH1–CH4 assigned as the positive input with the negative input being the analog ground (AGND) of the device. Finally, in the pseudo-differential mode CH1–CH3 are positive inputs referenced to CH4 which is now a pseudo-ground. This pseudo-ground input can be set to any potential within the input common-mode range of the converter. The analog signal conditioning required in transducer-based data acquisition systems is significantly simplified with this type of input flexibility. One converter package can now handle ground referenced inputs and true differential inputs as well as signals with some arbitrary reference voltage.

The analog input voltages for each channel can range from 50 mV below ground to 50 mV above V_{CC} (typically 5V) without degrading conversion accuracy.

2.0 REFERENCE CONSIDERATIONS

The voltage applied to the reference input of these converters defines the voltage span of the analog input (the difference between $V_{IN(MAX)}$ and $V_{IN(MIN)}$) over which the 256 possible output codes apply. The devices can be used in either ratiometric applications or in systems requiring absolute accuracy. The reference pin must be connected to a voltage source capable of driving the minimum reference input resistance of 1.1 kΩ. This pin is the top of a resistor divider string used for the successive approximation conversion.

In a ratiometric system (*Figure 2a*), the analog input voltage is proportional to the voltage used for the A/D reference. This voltage is typically the system power supply, so the V_{REF} pin can be tied to V_{CC}. This technique relaxes the stability requirements of the system reference as the analog input and A/D reference move together maintaining the same output code for a given input condition.

For absolute accuracy (*Figure 2b*), where the analog input varies between very specific voltage limits, the reference pin can be biased with a time and temperature stable voltage source. The LM385 and LM336 reference diodes are good low current devices to use with these converters.

The maximum value of the reference is limited to the V_{CC} supply voltage. The minimum value, however, can be quite small (see Typical Performance Characteristics) to allow direct conversions of transducer outputs providing less than a 5V output span. Particular care must be taken with regard to noise pickup, circuit layout and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter (1 LSB equals $V_{REF}/256$).

3.0 THE ANALOG INPUTS

3.1 Analog Differential Voltage Inputs and Common-Mode Rejection

The differential input of these converters actually reduces the effects of common-mode input noise, a signal common to both selected "+" and "−" inputs for a conversion (60 Hz is most typical). The time interval between sampling the

TABLE II. ADC0848 MUX Addressing

MUX Address					$\overline{CS}$	$\overline{WR}$	$\overline{RD}$	Channel									MUX Mode
MA4	MA3	MA2	MA1	MA0				CH1	CH2	CH3	CH4	CH5	CH6	CH7	CH8	AGND	
X	L	L	L	L	L	↓	H	+	−								Differential
X	L	L	L	H	L		H	−	+								
X	L	L	H	L	L		H			+	−						
X	L	L	H	H	L		H			−	+						
X	L	H	L	L	L		H					+	−				
X	L	H	L	H	L		H					−	+				
X	L	H	H	L	L		H							+	−		
X	L	H	H	H	L		H							−	+		
L	H	L	L	L	L	↓	H	+								−	Single-Ended
L	H	L	L	H	L		H		+							−	
L	H	L	H	L	L		H			+						−	
L	H	L	H	H	L		H				+					−	
L	H	H	L	L	L		H					+				−	
L	H	H	L	H	L		H						+			−	
L	H	H	H	L	L		H							+		−	
L	H	H	H	H	L		H								+	−	
H	H	L	L	L	L	↓	H	+							−		Pseudo-Differential
H	H	L	L	H	L		H		+						−		
H	H	L	H	L	L		H			+					−		
H	H	L	H	H	L		H				+				−		
H	H	H	L	L	L		H					+			−		
H	H	H	L	H	L		H						+		−		
H	H	H	H	L	L		H							+	−		
X	X	X	X	X	L	↓	L	Previous Channel Configuration										

Applications Information (Continued)

"+" input and then the "−" inputs is ½ of a clock period. The change in the common-mode voltage during this short time interval can cause conversion errors. For a sinusoidal common-mode signal this error is:

$$V_{ERROR(MAX)} = V_{peak}\,(2\pi\, f_{CM}) \times 0.5 \times \left(\frac{t_C}{8}\right)$$

where f_{CM} is the frequency of the common-mode signal, V_{peak} is its peak voltage value and t_C is the conversion time.

For a 60 Hz common-mode signal to generate a ¼ LSB error (≈5 mV) with the converter running at 40 μS, its peak value would have to be 5.43V. This large a common-mode signal is much greater than that generally found in a well designed data acquisition system.

3.2 Input Current

Due to the sampling nature of the analog inputs, short duration spikes of current enter the "+" input and exit the "−" input at the clock edges during the actual conversion. These currents decay rapidly and do not cause errors as the internal comparator is strobed at the end of a clock period. Bypass capacitors at the inputs will average these currents and cause an effective DC current to flow through the output resistance of the analog signal source. Bypass capacitors should not be used if the source resistance is greater than 1 kΩ.

3.3 Input Source Resistance

The limitation of the input source resistance due to the DC leakage currents of the input multiplexer is important. A worst-case leakage current of ± 1 μA over temperature will create a 1 mV input error with a 1 kΩ source resistance. An op amp RC active low pass filter can provide both impedance buffering and noise filtering should a high impedance signal source be required.

4.0 OPTIONAL ADJUSTMENTS

4.1 Zero Error

The zero of the A/D does not require adjustment. If the minimum analog input voltage value, $V_{IN(MIN)}$, is not ground, a zero offset can be done. The converter can be made to output 0000 0000 digital code for this minimum input voltage by biasing any V_{IN} (−) input at this $V_{IN(MIN)}$ value. This is useful for either differential or pseudo-differential modes of input channel configuration.

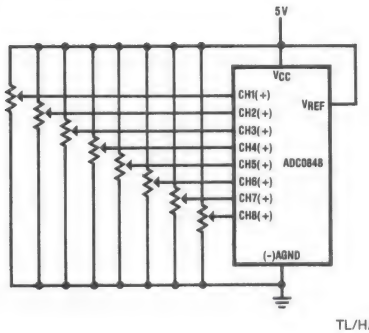
The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the V^- input and applying a small magnitude positive voltage to the V^+ input. Zero error is the difference between actual DC input voltage which is necessary to just cause an output digital code transition from 0000 0000 to 0000 0001 and the ideal ½ LSB value (½ LSB = 9.8 mV for $V_{REF} = 5.000\ V_{DC}$).

4.2 Full-Scale

The full-scale adjustment can be made by applying a differential input voltage which is 1 ½ LSB down from the desired analog full-scale voltage range and then adjusting the magnitude of the V_{REF} input for a digital output code changing from 1111 1110 to 1111 1111.

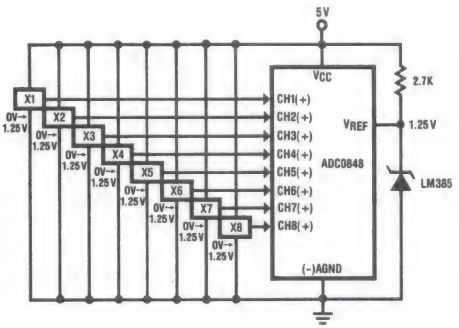
4.3 Adjusting for an Arbitrary Analog Input Voltage Range

If the analog zero voltage of the A/D is shifted away from ground (for example, to accommodate an analog input signal which does not go to ground), this new zero reference should be properly adjusted first. A V_{IN} (+) voltage which equals this desired zero reference plus ½ LSB (where the LSB is calculated for the desired analog span, 1 LSB = analog span/256) is applied to selected "+" input and the zero reference voltage at the corresponding "−" input should then be adjusted to just obtain the 00_{HEX} to 01_{HEX} code transition.



TL/H/5016–16

a) Ratiometric



TL/H/5016–17

b) Absolute with a Reduced Span

FIGURE 2. Referencing Examples

Applications Information (Continued)

The full-scale adjustment should be made [with the proper $V_{IN}(-)$ voltage applied] by forcing a voltage to the $V_{IN}(+)$ input which is given by:

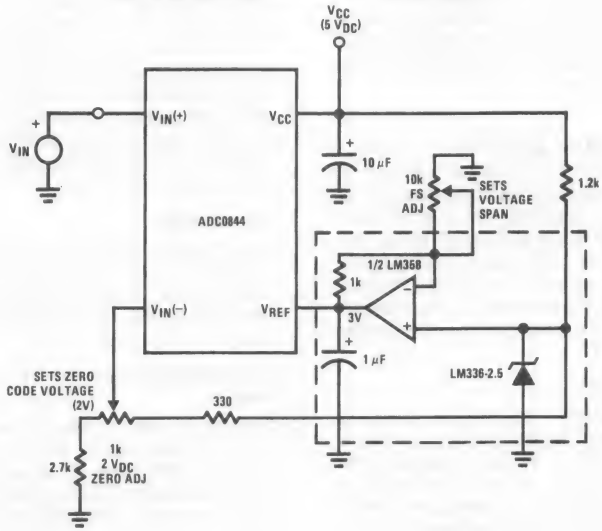
$$V_{IN}(+)\,\text{fs adj} = V_{MAX} - 1.5\left[\frac{(V_{MAX}-V_{MIN})}{256}\right]$$

where V_{MAX} = the high end of the analog input range and V_{MIN} = the low end (the offset zero) of the analog range. (Both are ground referenced.)

The V_{REF} (or V_{CC}) voltage is then adjusted to provide a code change from FE_{HEX} to FF_{HEX}. This completes the adjustment procedure.

For an example see the Zero-Shift and Span Adjust circuit below.

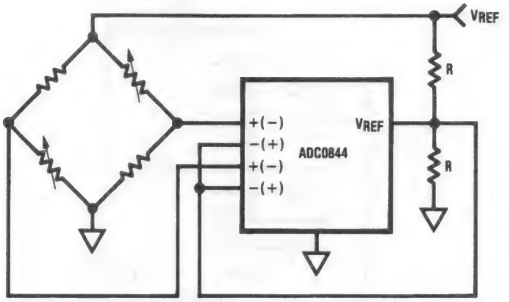
Zero-Shift and Span Adjust ($2V \leq V_{IN} \leq 5V$)



TL/H/5016–18

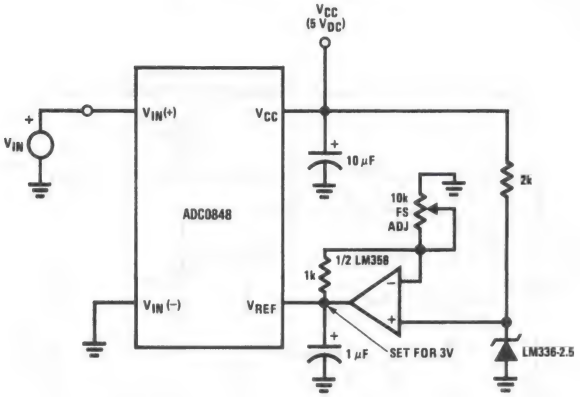
Applications Information (Continued)

Differential Voltage Input 9-Bit A/D



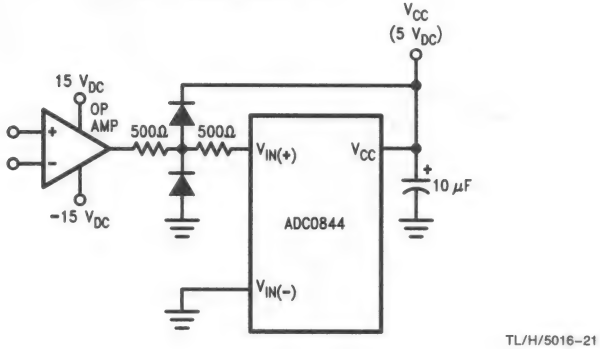
TL/H/5016–19

Span Adjust $0V \leq V_{IN} \leq 3V$



TL/H/5016–20

Protecting the Input

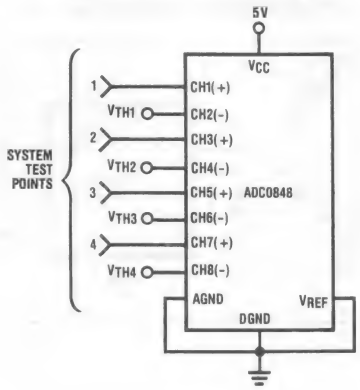


TL/H/5016–21

Diodes are 1N914

Applications Information (Continued)

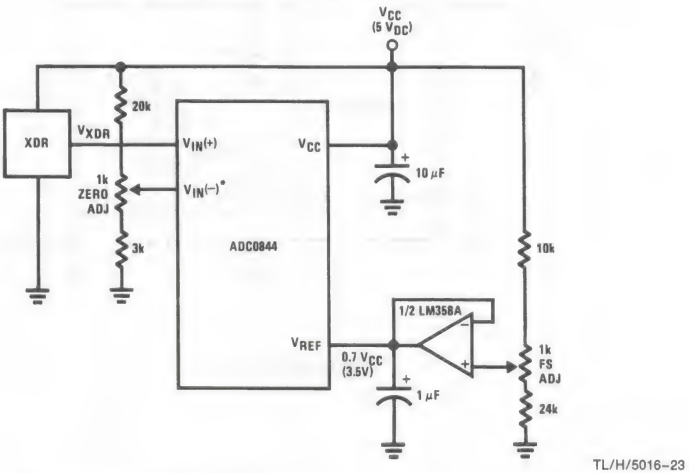
High Accuracy Comparators



TL/H/5016–22

DO = all 1s if $V_{IN}(+) > V_{IN}(-)$
DO = all 0s if $V_{IN}(+) < V_{IN}(-)$

Operating with Automotive Ratiometric Transducers

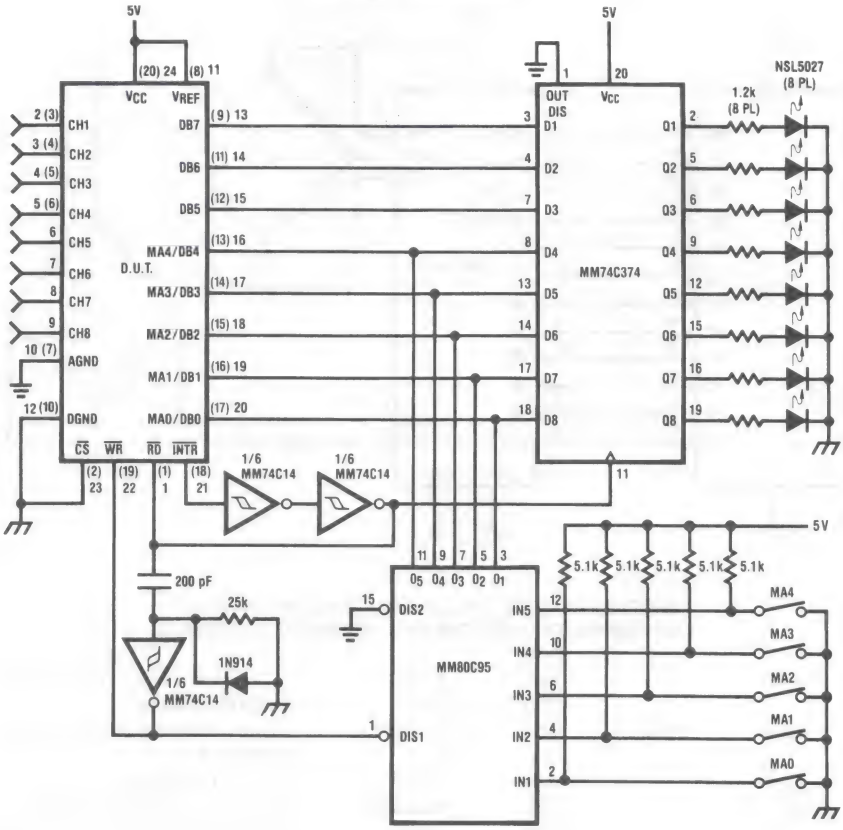


TL/H/5016–23

*$V_{IN}(-) = 0.15\ V_{CC}$
15% of $V_{CC} \leq V_{XDR} \leq 85\%$ of V_{CC}

Applications Information (Continued)

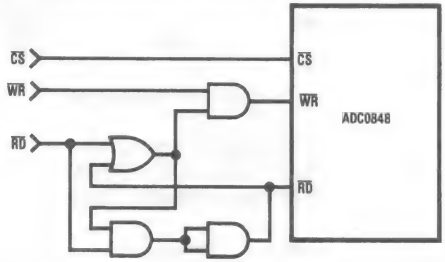
A Stand Alone Circuit



TL/H/5016–25

Note: DUT pin numbers in parentheses are for ADC0844, others are for ADC0848.

Start a Conversion without Updating the Channel Configuration



TL/H/5016–26

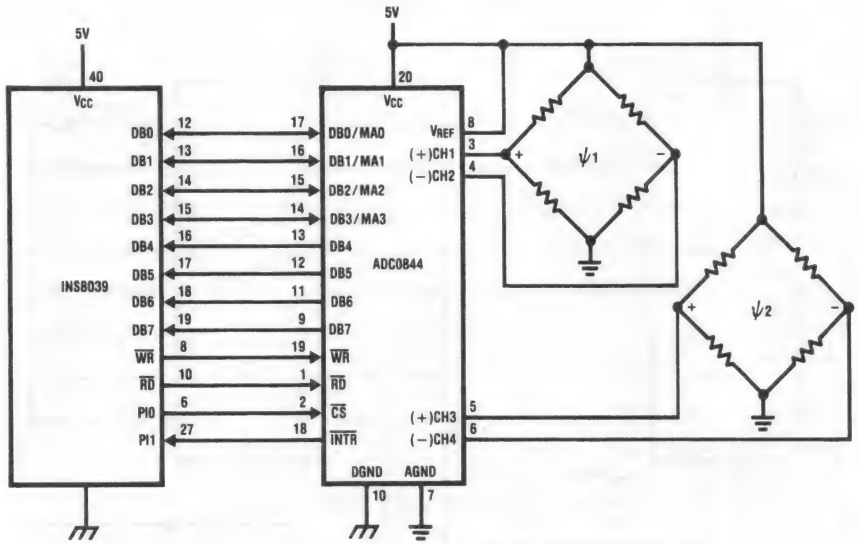
$\overline{CS}$•$\overline{WR}$ will update the channel configuration and start a conversion.

$\overline{CS}$•$\overline{RD}$ will read the conversion data and start a new conversion without updating the channel configuration.

Waiting for the end of this conversion is not necessary. A $\overline{CS}$•$\overline{WR}$ can immediately follow the $\overline{CS}$•$\overline{RD}$.

Applications Information (Continued)

ADC0844—INS8039 Interface



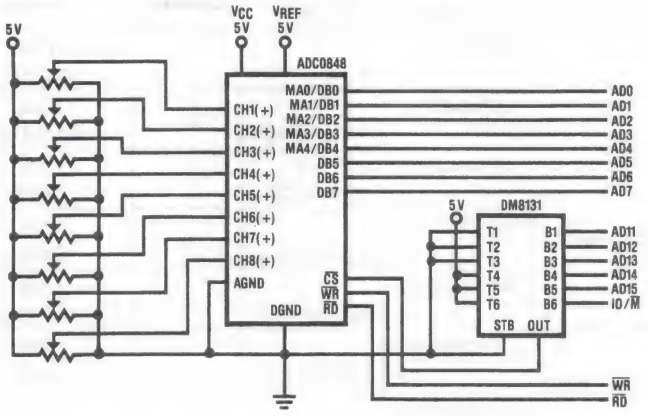
TL/H/5016–27

SAMPLE PROGRAM FOR ADC0844—INS8039 INTERFACE
CONVERTING TWO RATIOMETRIC, DIFFERENTIAL SIGNALS

```
                        ORG     0H
0000    04 10           JMP     BEGIN       ;START PROGRAM AT ADDR 10
                        ORG     10H         ;MAIN PROGRAM
0010    B9 FF   BEGIN:  MOV     R1,#0FFH    ;LOAD R1 WITH A UNUSED ADDR
                                            ;LOCATION
0012    B8 20           MOV     R0,#20H     ;A/D DATA ADDRESS
0014    89 FF           ORL     P1,#0FFH    ;SET PORT 1 OUTPUTS HIGH
0016    23 00           MOV     A,00H       ;LOAD THE ACC WITH A/D MUX DATA
                                            ;CH1 AND CH2 DIFFERENTIAL
0018    14 50           CALL    CONV        ;CALL THE CONVERSION SUBROUTINE
001A    23 02           MOV     A,#02H      ;LOAD THE ACC WITH A/D MUX DATA
                                            ;CH3 AND CH4 DIFFERENTIAL
001C    18              INC     R0          ;INCREMENT THE A/D DATA ADDRESS
001D    14 50           CALL    CONV        ;CALL THE CONVERSION SUBROUTINE

                        ;CONTINUE MAIN PROGRAM

                        ;CONVERSION SUBROUTINE
                        ;ENTRY:ACC—A/D MUX DATA
                        ;EXIT: ACC—CONVERTED DATA

                        ORG     50H
0050    99 FE   CONV:   ANL     P1,#0FEH    ;CHIP SELECT THE A/D
0052    91              MOVX    @R1,A       ;LOAD A/D MUX & START CONVERSION
0053    09      LOOP:   IN      A,P1        ;INPUT INTR STATE
0054    32 53           JB1     LOOP        ;IF INTR = 1 GOTO LOOP
0056    81              MOVX    A,@R1       ;IF INTR = 0 INPUT A/D DATA
0057    89 01           ORL     P1,&01H     ;CLEAR THE A/D CHIP SELECT
0059    A0              MOV     @R0,A       ;STORE THE A/D DATA
005A    83              RET                 ;RETURN TO MAIN PROGRAM
```

Applications Information (Continued)

I/O Interface to NSC800



TL/H/5016–28

SAMPLE PROGRAM FOR ADC0848—NSC800 INTERFACE

```
0008                    NCONV     EQU     16
000F                    DEL       EQU     15          ;DELAY 50 μsec CONVERSION
001F                    CS        EQU     1FH         ;THE BOARD ADDRESS
3C00                    ADDTA     EQU     003CH       ;START OF RAM FOR A/D
                                                      ;DATA
0000'  08 09 0A 0B      MUXDTA:   DB      08H,09H,0AH,0BH   ;MUX DATA
0004'  0C 0D 0E 0F                DB      0CH,0DH,0EH,0FH
0008'  0E 1F            START:    LD      C,CS
000A'  06 16                      LD      B,NCONV
000C'  21 0000'                   LD      HL,MUXDTA
000F'  11 003C                    LD      DE,ADDTA
0012'  ED A3            STCONV:   OUTI                ;LOAD A/D'S MUX DATA
                                                      ;AND START A CONVERSION
0014'  EB                         EX      DE,HL       ;HL = RAM ADDRESS FOR THE
                                                      ;A/D DATA
0015'  3E 0F                      LD      A,DEL
0017'  3D               WAIT:     DEC     A           ;WAIT 50 μsec FOR THE
0018'  C2 0013'                  JP      NZ,WAIT     ;CONVERSION TO FINISH
001B'  ED A2                      INI                 ;STORE THE A/D'S DATA
                                                      ;CONVERTED ALL INPUTS?
001D'  EB                         EX      DE,HL
001E'  C2 000E'                   JP      NZ,STCONV   ;IF NOT GOTO STCONV

                                  END
```

Note: This routine sequentially programs the MUX data latch in the signal-ended mode. For CH1-CH8 a conversion is started, then a 50 μs wait for the A/D to complete a conversion and the data is stored at address ADDTA for CH1, ADDTA + 1 for CH2, etc.

3

Ordering Information

Temperature Range	Total Unadjusted Error		MUX Channels	Package Outline
	±½ LSB	±1 LSB		
0°C to +70°C	ADC0844BCN	ADC0844CCN	4	N20A Molded Dip
	ADC0848BCN	ADC0848CCN	8	N24C Molded Dip
−40°C to +85°C	ADC0844BCJ	ADC0844CCJ	4	J20A Cerdip
	ADC0848BCJ	ADC0848CCJ	8	J24F Cerdip
	ADC0848BCV	ADC0848CCV	8	V28A Molded Chip Carrier
−55°C to +125°C	ADC0844BJ	ADC0844CJ	4	J20A Cerdip
	ADC0848BJ	ADC0848CJ	8	J24F Cerdip

National Semiconductor

ADC0852/ADC0854
Multiplexed Comparator with 8-Bit Reference Divider

General Description

The ADC0852 and ADC0854 are CMOS devices that combine a versatile analog input multiplexer, voltage comparator, and an 8-bit DAC which provides the comparator's threshold voltage (V_{TH}). The comparator provides a "1-bit" output as a result of a comparison between the analog input and the DAC's output. This allows for easy implementation of set-point, on-off or "bang-bang" control systems with several advantages over previous devices.

The ADC0854 has a 4 input multiplexer that can be software configured for single ended, pseudo-differential, and full-differential modes of operation. In addition the DAC's reference input is brought out to allow for reduction of the span.

The ADC0852 has a two input multiplexer that can be configured as 2 single-ended or 1 differential input pair. The DAC reference input is internally tied to V_{CC}.

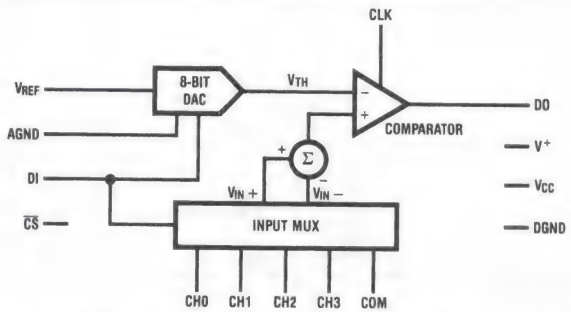
The multiplexer and 8-bit DAC are programmed via a serial data input word. Once programmed the output is updated once each clock cycle up to a maximum clock rate of 400 kHz.

Features

- 2 or 4 channel multiplexer
- Differential or Single-ended input, software controlled
- Serial digital data interface
- 256 programmable reference voltage levels
- Continuous comparison after programming
- Fixed, ratiometric, or reduced span reference capability (ADC 0854)

Key Specifications

- Accuracy, ± ½ LSB or ± 1 LSB of Reference (0.2%)
- Single 5V power supply
- Low Power, 15 mW



TL/H/5521–1

FIGURE 1. ADC0854 Simplified Block Diagram (ADC0852 has 2 input channels, COM tied to GND, V_{REF} tied to V_{CC}, V+ omitted, and one GND connection)

2 Channel and 4 Channel Pin Out

ADC0852 2-CHANNEL MUX
Dual-In-Line Package

TL/H/5521–10

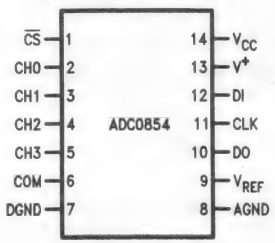
Top View

AGND and COM internally connected to GND

V_{REF} internally connected to V_{CC}

Order Number ADC0852
See NS Package Number J08A or N08E

ADC0854 4-CHANNEL MUX
Dual-In-Line Package



TL/H/5521–11

Top View

Order Number ADC0854
See NS Package Number J14A or N14A

3

Absolute Maximum Ratings (Notes 1 and 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Current into V^+ (Note 3)	15 mA
Supply Voltage, V_{CC} (Note 3)	6.5V
Voltage	
Logic and Analog Inputs	$-0.3V$ to $V_{CC} + 0.3V$
Input Current per Pin	± 5 mA
Input Current per Package	± 20 mA
Storage Temperature	$-65°C$ to $+150°C$
Package Dissipation at $T_A = 25°C$ (Board Mount)	0.8W
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
ESD Susceptibility (Note 14)	2000V

Operating Conditions

Supply Voltage, V_{CC}	$4.5V_{DC}$ to $6.3V_{DC}$
Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC0854BJ, ADC0854CJ	$-55°C \le T_A \le 125°C$
ADC0852BJ, ADC0852CJ	
ADC0854BCJ, ADC0854CCJ	$-40°C \le T_A \le 85°C$
ADC0852BCJ, ADC0852CCJ	
ADC0854BCN, ADC0854CCN	$0°C \le T_A \le 70°C$
ADC0852BCN, ADC0852CCN	

Electrical Characteristics

The following specifications apply for $V_{CC} = V^+ = 5V$ (no V^+ on ADC0852), $V_{REF} \le V_{CC} + 0.1V$, $f_{CLK} = 250$ kHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Parameter	Conditions	ADC0852BCJ/CCJ/BJ/CJ ADC0854BCJ/CCJ/BJ/CJ			ADC0852BCN/CCN ADC0854BCN/CCN			Units
		Typ (Note 4)	Tested Limit (Note 5)	Design Limit (Note 6)	Typ (Note 4)	Tested Limit (Note 5)	Design Limit (Note 6)	
CONVERTER AND MULTIPLEXER CHARACTERISTICS								
Total Unadjusted Error (Note 7)	V_{REF} Forced to 5.000 V_{DC}							
ADC0852/4/BCN						$\pm$½	**$\pm$½**	LSB
ADC0852/4/BJ/BCJ			**$\pm$½**					LSB
ADC0852/4/CCN						$\pm$1	**$\pm$1**	LSB
ADC0852/4/CJ/CCJ			**$\pm$1**					LSB
Comparator Offset								
ADC0852/4/BCN		2.5			2.5		**10**	mV
ADC0852/4/BJ/BCJ		2.5	**10**		2.5			mV
ADC0852/4/CCN		2.5			2.5		**20**	mV
ADC0852/4/CCJ		2.5	**20**		2.5			mV
Minimum Total Ladder Resistance	ADC0854 (Note 15)	3.5	**1.3**		3.5	1.3	**1.3**	kΩ
Maximum Total Ladder Resistance	ADC0854 (Note 15)	3.5	**5.9**		3.5	5.4	**5.9**	kΩ
Minimum Common-Mode Input (Note 8)	All MUX Inputs and COM Input		**GND−0.05**			GND−0.05	**GND−0.05**	V
Maximum Common-Mode Input (Note 8)	All MUX Inputs and COM Input		**V_{CC} + 0.05**			V_{CC} + 0.05	**V_{CC} + 0.05**	V
DC Common-Mode Error		$\pm$1/16	$\pm$¼		$\pm$1/16	$\pm$¼	$\pm$¼	LSB
Power Supply Sensitivity	$V_{CC} = 5V \pm 5\%$	$\pm$1/16	$\pm$¼		$\pm$1/16	$\pm$¼	$\pm$¼	LSB
V_Z, Internal diode breakdown at V^+ (Note 3) MIN	15 mA into V^+		**6.3**			6.3		V
MAX			**8.5**			8.5		V
I_{OFF}, Off Channel Leakage Current (Note 9)	On Channel = 5V, Off Channel = 0V		**−1**				**−1**	µA
			−200			−200		nA
	On Channel = 0V, Off Channel = 5V		**+1**				**+1**	µA
			+200			+200		nA

Electrical Characteristics (Continued)

The following specifications apply for $V_{CC} = V^+ = 5V$ (no V^+ on ADC0852), $f_{CLK} = 250$ kHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$.

Parameter	Conditions	ADC0852BCJ/CCJ/BJ/CJ ADC0854BCJ/CCJ/BJ/CJ			ADC0852BCN/CCN ADC0854BCN/CCN			Units
		Typ (Note 4)	Tested Limit (Note 5)	Design Limit (Note 6)	Typ (Note 4)	Tested Limit (Note 5)	Design Limit (Note 6)	
CONVERTER AND MULTIPLEXER CHARACTERISTICS (Continued)								
I_{ON}, On Channel Leakage Current (Note 9)	On Channel = 5V, Off Channel = 0V		**+1** +200			+200	**+1**	μA nA
	On Channel = 0V, Off Channel = 5V		**−1** −200			−200	**−1**	μA nA
DIGITAL AND DC CHARACTERISTICS								
$V_{IN(1)}$, Logical "1" Input Voltage	$V_{CC} = 5.25V$		**2.0**			2.0	**2.0**	V
$V_{IN(0)}$, Logical "0" Input Voltage	$V_{CC} = 4.75V$		**0.8**			0.8	**0.8**	V
$I_{IN(1)}$, Logical "1" Input Current	$V_{IN} = V_{CC}$	0.005	**1**		0.005	1	**1**	μA
$I_{IN(0)}$, Logical "0" Input Current	$V_{IN} = 0V$	−0.005	**−1**		−0.005	−1	**−1**	μA
$V_{OUT(1)}$, Logical "1" Output Voltage	$V_{CC} = 4.75V$ $I_{OUT} = -360\ \mu A$ $I_{OUT} = -10\ \mu A$		 **2.4** **4.5**			 2.4 4.5	 **2.4** **4.5**	 V V
$V_{OUT(0)}$, Logical "0" Output Voltage	$I_{OUT} = 1.6$ mA, $V_{CC} = 4.75V$		**0.4**			0.4	**0.4**	V
I_{OUT}, TRI-STATE® Output Current (DO)	$\overline{CS}$ = Logical "1" $V_{OUT} = 0.4V$ $V_{OUT} = 5V$	 −0.1 0.1	 **−3** **3**		 −0.1 0.1	 −3 3	 **−3** **3**	 μA μA
I_{SOURCE}	V_{OUT} Short to GND	−14	**−6.5**		−14	−7.5	**−6.5**	mA
I_{SINK}	V_{OUT} Short to V_{CC}	16	**8.0**		16	9.0	**8.0**	mA
I_{CC} Supply Current ADC0852	Includes DAC Ladder Current	2.7	**6.5**		2.7	6.5	**6.5**	mA
I_{CC} Supply Current ADC0854 (Note 3)	Does not Include DAC Ladder Current	0.9	**2.5**		0.9	2.5	**2.5**	mA

AC Characteristics $t_r = t_f = 20$ ns, $T_A = 25°C$

Symbol	Parameter		Conditions	Typ (Note 4)	Tested Limit (Note 5)	Design Limit (Note 6)	Units
f_{CLK}	Clock Frequency (Note 12)	MIN MAX			10	 400	kHz kHz
t_{D1}	Rising Edge of Clock to "DO" Enabled		C_L = 100 pF	650		1000	ns
t_r	Comparator Response Time (Note 13)		Not Including Addressing Time			2 + 1 μs	$1/f_{CLK}$
	Clock Duty Cycle (Note 10)	MIN MAX			40 60		% %
t_{SET-UP}	CS Falling Edge or Data Input Valid to CLK Rising Edge	MAX				250	ns
t_{HOLD}	Data Input Valid after CLK Rising Edge	MIN				90	ns
t_{pd1}, t_{pd0}	CLK Falling Edge to Output Data Valid (Note 11)	MAX	C_L = 100 pF	650		1000	ns
t_{1H}, t_{0H}	Rising Edge of CS to Data Output Hi-Z	MAX	C_L = 10 pF, R_L = 10k C_L = 100 pF, R_L = 2k (see TRI-STATE Test Circuits)	125	 500	250 500	ns ns
C_{IN}	Capacitance of Logic Input			5			pF
C_{OUT}	Capacitance of Logic Outputs			5			pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to ground.

Note 3: Internal zener diodes (approx. 7V) are connected from V+ to GND and V_{CC} to GND. The zener at V+ can operate as a shunt regulator and is connected to V_{CC} via a conventional diode. Since the zener voltage equals the A/D's breakdown voltage, the diode ensures that V_{CC} will be below breakdown when the device is powered from V+. Functionality is therefore guaranteed for V+ operation even though the resultant voltage at V_{CC} may exceed the specified Absolute Max of 6.5V. It is recommended that a resistor be used to limit the max current into V+.

Note 4: Typicals are at 25°C and represent most likely parametric norm.

Note 5: Tested and guaranteed to National AOQL (Average Outgoing Quality Level).

Note 6: Guaranteed, but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Note 7: Total unadjusted error includes comparator offset, DAC linearity, and multiplexer error. It is expressed in LSBs of the threshold DAC's input code.

Note 8: For $V_{IN}(-) \geq V_{IN}(+)$ the output will be 0. Two on-chip diodes are tied to each analog input (see Block Diagram) which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct—especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

Note 9: Leakage current is measured with the clock not switching.

Note 10: A 40% to 60% clock duty cycle range ensures proper operation at all clock frequencies. In the case that an available clock has a duty cycle outside of these limits then 1.6 μS ≤ CLK Low ≤ 60 μS and 1.6 μS ≤ CLK HIGH ≤ ∞.

Note 11: With $\overline{CS}$ low and programming complete, D0 is updated on each falling CLK edge. However, each new output is based on the comparison completed 0.5 clock cycles prior (see *Figure 5*).

Note 12: Error specs are not guaranteed at 400 kHz (see graph: Comparator Error vs. f_{CLK}).

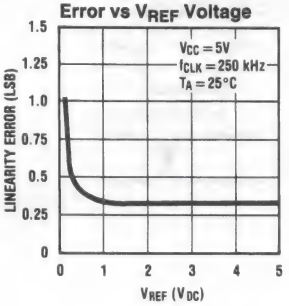
Note 13: See text, section 1.2.

Note 14: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

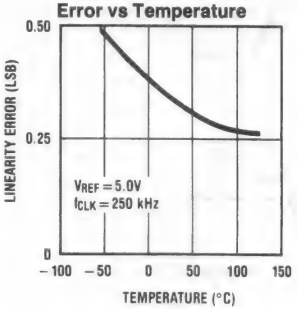
Note 15: Because the reference ladder of the ADC0852 is internally connected to V_{CC}, ladder resistance cannot be directly tested for the ADC0852. Ladder current is included in the ADC0852's supply current specification.

Typical Performance Characteristics

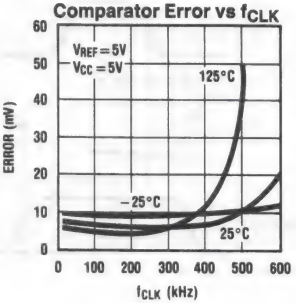
Internal DAC Linearity Error vs V_{REF} Voltage



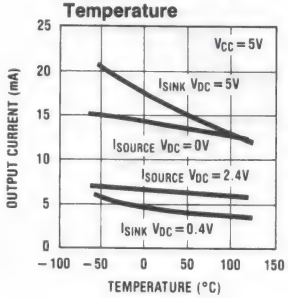
Internal DAC Linearity Error vs Temperature



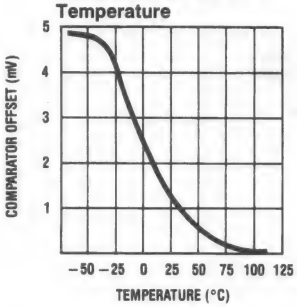
Comparator Error vs f_{CLK}



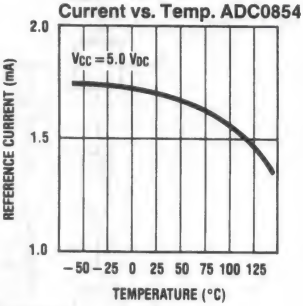
Output Current vs Temperature



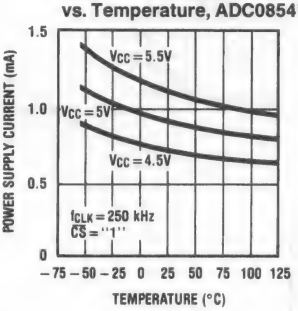
Comparator Offset vs Temperature



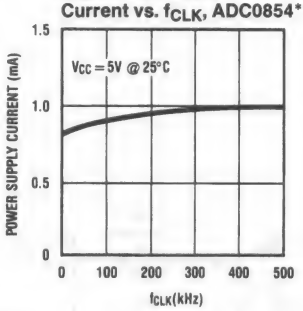
I_{REF}, Reference Current vs. Temp. ADC0854



I_{CC}, Power Supply Current vs. Temperature, ADC0854*



I_{CC}, Power Supply Current vs. f_{CLK}, ADC0854*



*For ADC0852 add I_{REF}

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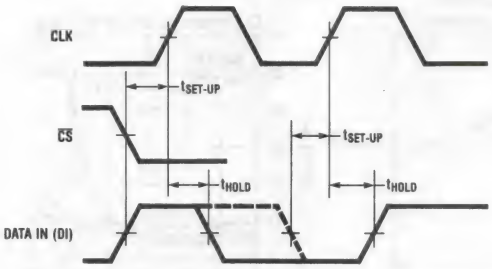
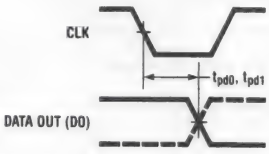
Timing Diagrams

Data Input Timing



TL/H/5521–3

Data Output Timing



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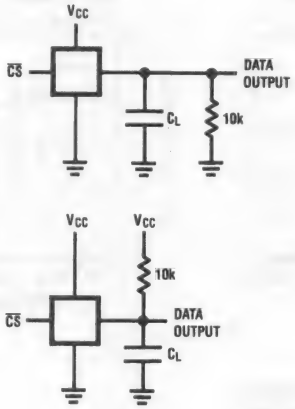
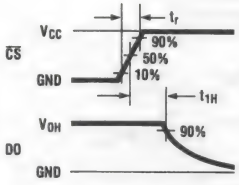
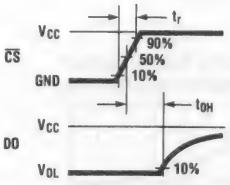
TRI-STATE Test Circuits and Waveforms

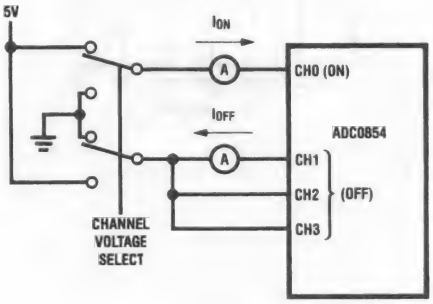




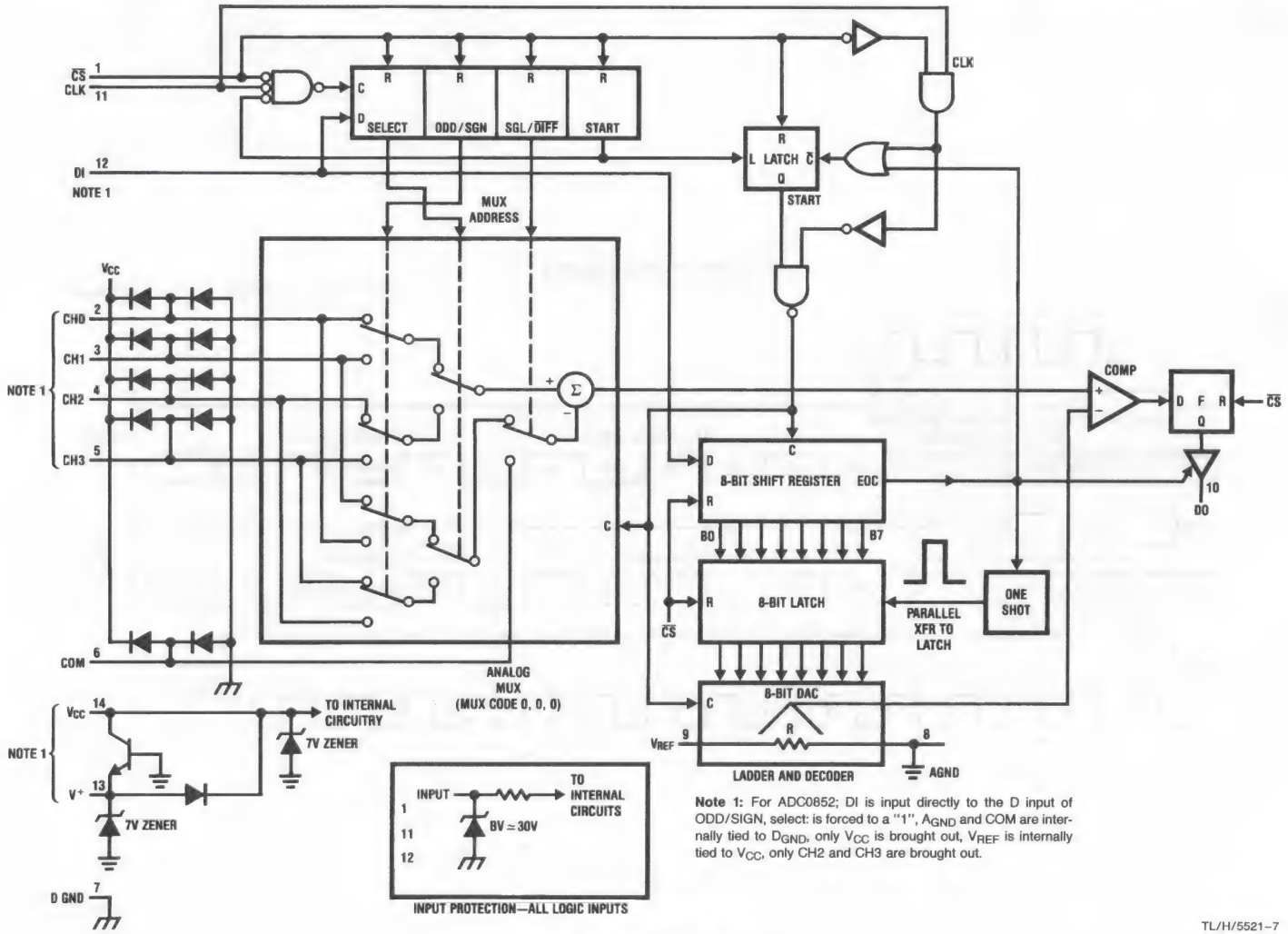


TL/H/5521–5

Leakage Test Circuit



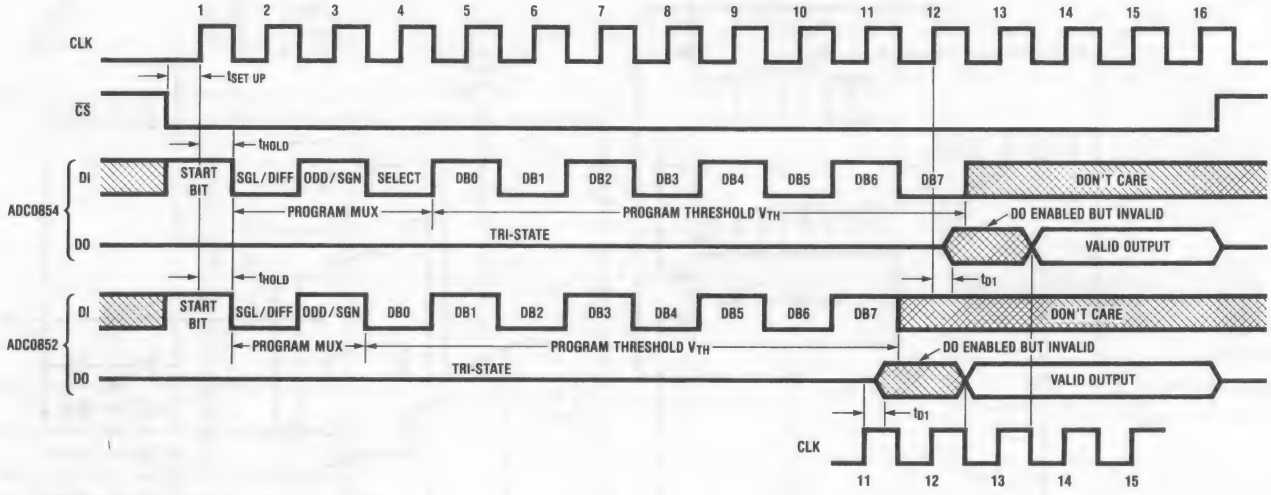
TL/H/5521–6



Note 1: For ADC0852; DI is input directly to the D input of ODD/SIGN, select: is forced to a "1", A_{GND} and COM are internally tied to D_{GND}, only V_{CC} is brought out, V_{REF} is internally tied to V_{CC}, only CH2 and CH3 are brought out.

TL/H/5521–7

FIGURE 2. Detailed Block Diagram



TL/H/5521–12

Note: Valid Output can change only on Falling Edge of CLK.

FIGURE 3. Timing Diagram

Functional Description

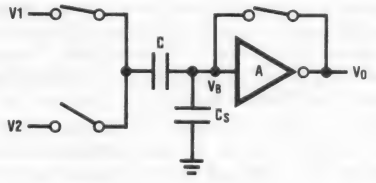
1. 1 The Sampled-data Comparator

The ADC0852 and ADC0854 utilize a sampled-data comparator structure to compare the analog difference between a selected "+" and "−" input to an 8-bit programmable threshold.

This comparator consists of a CMOS inverter with a capacitively coupled input (*Figure 4*). Analog switches connect the two comparator inputs to the input capacitor and also connect the inverter's input and output. This device in effect now has one differential input pair. A comparison requires two cycles, one for zeroing the comparator and another for making the comparison.

In the first cycle (*Figure 4a*), one input switch and the inverter's feedback switch are closed. In this interval, the input capacitor (C) is charged to the connected input (V1) less the inverter's bias voltage (V_B, approx. 1.2 volts). In the second cycle (*Figure 4b*) these two switches are opened and the other (V2) input's switch is closed. The input capacitor now subtracts its stored voltage from the second input and the difference is amplified by the inverter's open loop gain. The inverter input (V_B') becomes $V_B - (V1 - V2)\frac{C}{C + C_S}$ and the output will go high or low depending on the sign of V_B'-V_B.

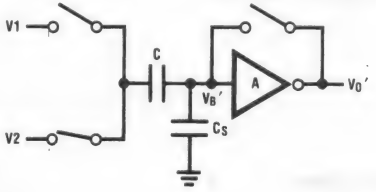
FIGURE 4. Sampled-Data Comparator



- $V_0 = V_B$
- V on C = $V_1 - V_B$
- C_S = Stray Input Node Cap.
- V_B = Inverter Input Bias Voltage

TL/H/5521–8

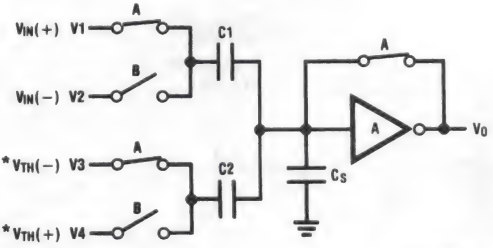
FIGURE 4a. Zeroing Phase



- $V_{B'} - V_B = (V_2 - V_1)\frac{C}{C + C_S}$
- $V_0 = \frac{-A}{C + C_S}[CV_2 - CV_1]$
- V_0 is dependent on $V_2 - V_1$

TL/H/5521–9

FIGURE 4b. Compare Phase



$$V_0 = \frac{-A}{C_1 + C_2 + C_S}[C_1(V_2 - V_1) + C_2(V_4 - V_3)]$$

$$= \frac{-A}{C_1 + C_2 + C_S}[\Delta Q C_1 + \Delta Q C_2]$$

* Comparator Reads V_{TH} from Internal DAC Differentially

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FIGURE 4c. Multiple Differential Inputs

Functional Description (Continued)

In actual practice, the devices used in the ADC0852/4 are a simple but important expansion of the basic comparator described above. As shown in *Figure 4c*, multiple differential comparisons can be made. In this circuit, the feedback switch and one input switch on each capacitor (A switches) are closed in the first cycle. Then the other input on each capacitor is connected while all of the first switches are opened. The change in voltage at the inverter's input, as a result of the change in charge on each input capacitor (C1, C2), will now depend on both input signal differences.

1.2 Input Sampling and Response Time

The input phases of the comparator relate to the device clock (CLK) as shown in *Figure 5*. Because the comparator is a sampling device, its response characteristics are somewhat different from those of linear comparators. The $V_{IN}(+)$ input is sampled first (CLK high) followed by $V_{IN}(-)$ (CLK low). The output responds to those inputs, one half cycle later, on CLK's falling edge.

The comparator's response time to an input step is dependent on the step's phase relation to the CLK signal. If an input step occurs too late to influence the most imminent comparator decision, one more CLK cycle will pass before the output is correct. In effect, the response time for the $V_{IN}(+)$ input has a minimum of 1 CLK cycle + 1 μS and a maximum of 2 CLK cycles + 1 μS. The $V_{IN}(-)$ input's delay will range from 1/2 CLK cycle + 1 μS to 1.5 CLK cycles + 1 μS since it is sampled after $V_{IN}(+)$.

The sampled inputs also affect the device's response to pulsed signals. As shown in the shaded areas in *Figure 5*, pulses that rise and/or fall near the latter part of a CLK half-cycle may be ignored.

1.3 Input Multiplexer

A unique input multiplexing scheme has been utilized to provide multiple analog channels with software-configurable single-ended, differential, or pseudo-differential operation. The analog signal conditioning required in transducer-input and other types of data acquisition systems is significantly simplified with this type of input flexibility. One device package can now handle ground referenced inputs as well as signals with some arbitrary reference voltage.

On the ADC0854, the "common" pin (pin 6) is used as the "−" input for all channels in single-ended mode. Since this input need not be at analog ground, it can be used as the common line for pseudo-differential operation. It may be tied to a reference potential that is common to all inputs and within the input range of the comparator. This feature is especially useful in single-supply applications where the analog circuitry is biased to a potential other than ground.

A particular input configuration is assigned during the MUX addressing sequence which occurs prior to the start of a comparison. The MUX address selects which of the analog channels is to be enabled, what the input mode will be, and the input channel polarity. One limitation is that differential inputs are restricted to adjacent channel pairs. For example, channel 0 and 1 may be selected as a differential pair but they cannot act differentially with any other channel.

The channel and polarity selection is done serially via the DI input. A complete listing of the input configurations and corresponding MUX addresses for the ADC0852 and ADC0854 is shown in tables I and II. *Figure 6* illustrates the analog connections for the various input options.

The analog input voltage for each channel can range from 50 mV below ground to 50 mV above V_{CC} (typically 5V) without degrading accuracy.

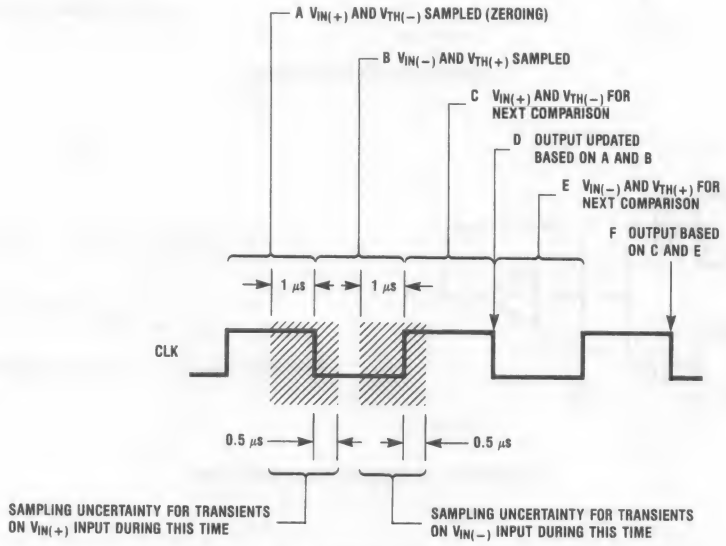
A VIN(+) AND VTH(−) SAMPLED (ZEROING)
B VIN(−) AND VTH(+) SAMPLED
C VIN(+) AND VTH(−) FOR NEXT COMPARISON
D OUTPUT UPDATED BASED ON A AND B
E VIN(−) AND VTH(+) FOR NEXT COMPARISON
F OUTPUT BASED ON C AND E
1 µs
1 µs
CLK
0.5 µs
0.5 µs
SAMPLING UNCERTAINTY FOR TRANSIENTS ON VIN(+) INPUT DURING THIS TIME
SAMPLING UNCERTAINTY FOR TRANSIENTS ON VIN(−) INPUT DURING THIS TIME

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FIGURE 5. Analog Input Timing

Functional Description (Continued)

TABLE I. MUX Addressing: ADC0854 Single-Ended MUX Mode

MUX Address			Channel				
SGL/$\overline{\text{DIF}}$	ODD/SIGN	SELECT	0	1	2	3	COM
1	0	0	+				−
1	0	1			+		−
1	1	0		+			−
1	1	1				+	−

Differential MUX Mode

MUX Address			Channel			
SGL/$\overline{\text{DIF}}$	ODD/SIGN	SELECT	0	1	2	3
0	0	0	+	−		
0	0	1			+	−
0	1	0	−	+		
0	1	1			−	+

TABLE II. MUX Addressing: ADC0852 Single Ended MUX Mode

MUX Address		Channel	
SGL/$\overline{\text{DIF}}$	ODD/SIGN	0	1
1	0	+	
1	1		+

COM is internally tied to A GND

Differential MUX Mode

MUX Address		Channel	
SGL/$\overline{\text{DIF}}$	ODD/SIGN	0	1
0	0	+	−
0	1	−	+

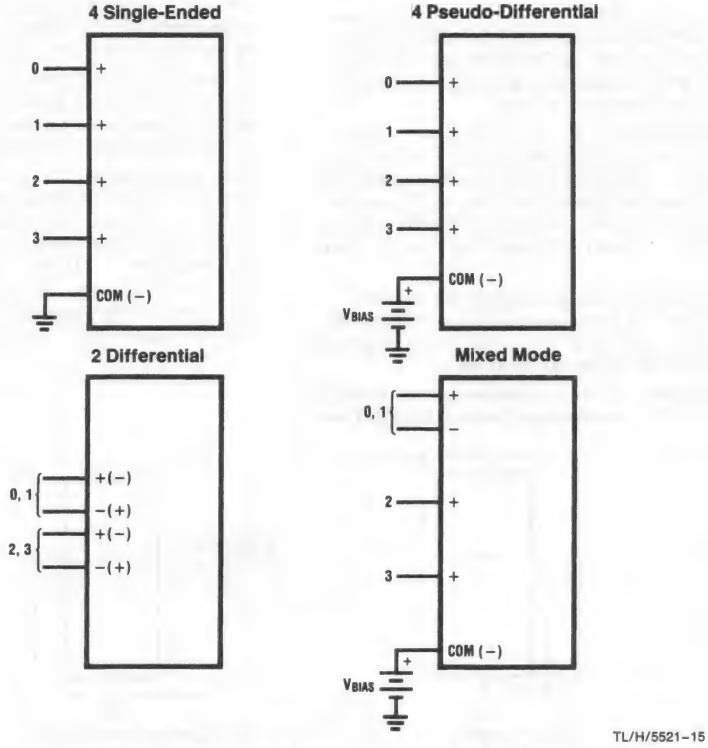

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FIGURE 6. Analog Input Multiplexer Options for the ADC0854

Functional Description (Continued)

2.0 THE DIGITAL INTERFACE

An important characteristic of the ADC0852 and ADC0854 is their serial data link with the controlling processor. A serial communication format eliminates the transmission of low level analog signals by locating the comparator close to the signal source. Thus only highly noise immune digital signals need to be transmitted back to the host processor.

To understand the operation of these devices it is best to refer to the timing diagrams (*Figure 3*) and functional block diagram (*Figure 2*) while following a complete comparison sequence.

1. A comparison is initiated by first pulling the $\overline{CS}$ (chip select) line low. This line must be held low for the entire addressing sequence and comparison. The comparator then waits for a start bit, its MUX assignment word, and an 8-bit code to set the internal DAC which supplies the comparator's threshold voltage (V_{TH}).

2. An external clock is applied to the CLK input. This clock can be applied continuously and need not be gated on and off.

3. On each rising edge of the clock, the level present on the DI line is clocked into the MUX address shift register. The start bit is the first logic "1" that appears on this line. All leading zeroes are ignored. After the start bit, the ADC0852 expects the next 2 bits to be the MUX assignment word while the ADC0854, with more MUX configurations, looks for 3 bits.

4. Immediately after the MUX assignment word has been clocked in, the shift register then reads the next eight bits as the input code to the internal DAC. This eight bit word is read LSB first and is used to set the voltage applied to the comparator's threshold input (internal).

5. After the rising edge of the 11th or 12th clock (ADC0852 or ADC0854 respectively) following the start bit, the comparator and DAC programming is complete. At this point the DI line is disabled and ignores further inputs. Also at this time the data out (DO) line comes out of TRI-STATE and enters a don't care state (undefined output) for 1.5 clock cycles.

6. The result of the comparison between the programmed threshold voltage and the difference between the two selected inputs ($V_{IN}(+) - V_{IN}(-)$) is output to the DO line on each subsequent high to low clock transition.

7. After programming, continuous comparison on the same selected channel with the same programmed threshold can be done indefinitely, without reprogramming the device, as long as $\overline{CS}$ remains low. Each new comparator decision will be shifted to the output on the falling edge of the clock. However, the output will, in effect, "lag" the analog input by 0.5 to 1.5 clock cycles because of the time required to make the comparison and latch the output (see *Figure 5*).

8. All internal registers are cleared when the $\overline{CS}$ line is brought high. If another comparison is desired $\overline{CS}$ must make a high to low transition followed by new address and threshold programming.

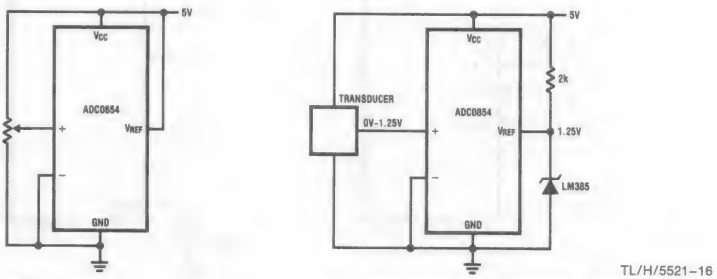
3.0 REFERENCE CONSIDERATIONS / RATIOMETRIC OPERATION

The voltage applied to the "V_{REF}" input of the DAC defines the voltage span that can be programmed to appear at the threshold input of the comparator. The ADC0854 can be used in either ratiometric applications or in systems with absolute references. The V_{REF} pin must be connected to a source capable of driving the DAC ladder resistance (typ. 2.4 kΩ) with a stable voltage.

In ratiometric systems, the analog input voltage is normally a proportion of the DAC's or A/D's reference voltage. For example, a mechanical position servo using a potentiometer to indicate rotation, could use the same voltage to drive the reference as well as the potentiometer. Changes in the value of V_{REF} would not affect system accuracy since only the relative value of these signals to each other is important. This technique relaxes the stability requirements of the system reference since the analog input and DAC reference move together, thus maintaining the same comparator output for a given input condition.

In the absolute case, the V_{REF} input can be driven with a stable voltage source whose output is insensitive to time and temperature changes. The LM385 and LM336 are good low current devices for this purpose.

The maximum value of V_{REF} is limited to the V_{CC} supply voltage. The minimum value can be quite small (see typical performance curves) allowing the effective resolution of the comparator threshold DAC to also be small (V_{REF} = 0.5V, DAC resolution = 2.0 mV). This in turn lets the designer have finer control over the comparator trip point. In such instances however, more care must be taken with regard to noise pickup, grounding, and system error sources.



a) Ratiometric

b) Absolute with a Reduced Span

FIGURE 7. Referencing Examples

Functional Description (Continued)

4.0 ANALOG INPUTS

4. 1 Differential Inputs

The serial interface of the ADC0852 and ADC0854 allows them to be located right at the analog signal source and to communicate with a controlling processor via a few fairly noise immune digital lines. This feature in itself greatly reduces the analog front end circuitry often needed to maintain signal integrity. Nevertheless, a few words are in order with regard to the analog inputs should the input be noisy to begin with or possibly riding on a large common mode voltage.

The differential input of the comparator actually reduces the effect of common-mode input noise, i.e. signals common to both selected "+" and "−" inputs such as 60 Hz line noise. The time interval between sampling the "+" input and then the "−" input is ½ of a clock period (see *Figure 5).*

The change in the common-mode voltage during this short time interval can cause comparator errors. For a sinusoidal common-mode signal this error is:

V_{ERROR} (MAX) = V_{PEAK} (2π f_{CM}/2 f_{CLK})

where f_{CM} is the frequency of the common-mode signal, V_{peak} is its peak voltage value, and f_{CLK} is the DAC clock frequency.

For example, 1 V_{PP} 60 Hz noise superimposed on both sides of a differential input signal would cause an error (referred to the input) of 0.75 mV. This amounts to less than 1/25 of an LSB referred to the threshold DAC, (assuming V_{REF} = 5V and f_{CLK} = 250 kHz).

4. 2 Input Currents and Filtering

Due to the sampling nature of the analog inputs, short spikes of current enter the "+" input and leave the "−" at the clock edges during a comparison. These currents decay rapidly and do not cause errors as the comparator is strobed at the end of the clock period (see *Figure 5*).

The source resistance of the analog input is important with regard to the DC leakage currents of the input multiplexer. The worst-case leakage currents of ±1 μA over temperature will create a 1 mV input error with a 1 kΩ source resistance. An op-amp RC active low pass filter can provide both impedance buffering and noise filtering should a high impedance source be required.

4. 3 Arbitrary Analog Input/Reference Range

The total span of the DAC output and hence the comparator's threshold voltage is determined by the DAC reference. For example, if V_{REF} is set to 1 volt then the comparator's threshold can be programmed over a 0 to 1 volt range with 8 bits of resolution. From the analog input's point of view, this span can also be shifted by applying an offset potential to one of the comparator's selected analog input lines (usually "−"). This gives the designer greater control of the ADC0852/4's input range and resolution and can help simplify or eliminate expensive signal conditioning electronics.

An example of this capability is shown in the "Load Cell Limit Comparator" of *Figure 15.* In this circuit, the ADC0852 allows the load-cell signal conditioning to be done with only one dual op-amp and without complex, multiple resistor matching.

5.0 POWER SUPPLY

A unique feature of the ADC0854 is the inclusion of a 7 volt zener diode connected from the "V+" terminal to ground (*Figures 2* and *8*) "V+" also connects to "V_{CC}" via a silicon diode. The zener is intended for use as a shunt voltage regulator to eliminate the need for additional regulating components. This is especially useful if the ADC0854 is to be remotely located from the system power source.

An important use of the interconnecting diode between V+ and V_{CC} is shown in *Figures 10* and *11.* Here this diode is used as a rectifier to allow the V_{CC} supply for the converter to be derived from the comparator clock. The low device current requirements and the relatively high clock frequencies used (10 kHz–400 kHz) allows use of the small value filter capacitor shown. The shunt zener regulator can also be used in this mode however this requires a clock voltage swing in excess of 7 volts. Current limiting for the zener is also needed, either built into the clock generator or through a resistor connected from the clock to V+.

Typical Applications

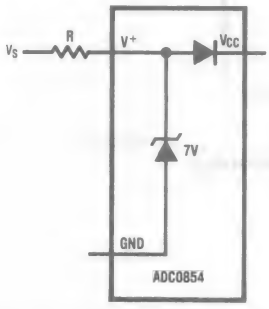


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FIGURE 8. An On-Chip Shunt Regulator Diode

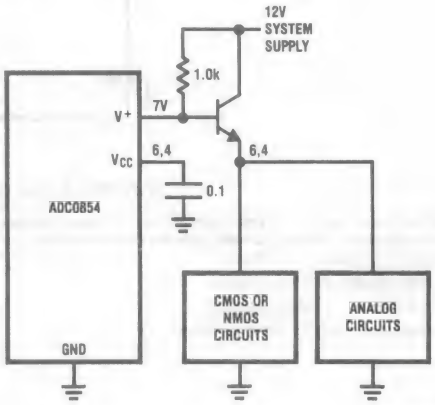

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FIGURE 9. Using the ADC0854 as the System Supply Regulator

Typical Applications (Continued)

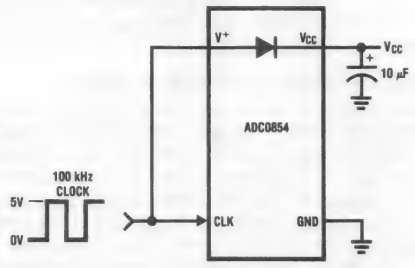

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FIGURE 10. Generating V_{CC} from the Comparator Clock

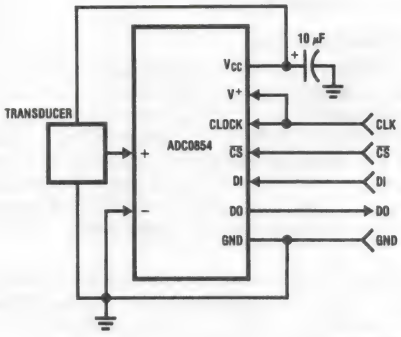

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FIGURE 11. Remote Sensing—Clock and Power on One Wire

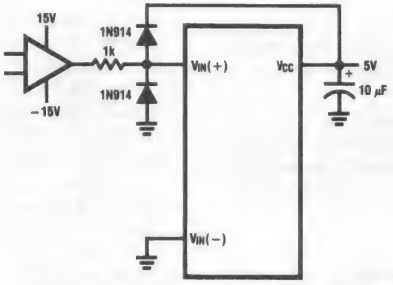

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FIGURE 12. Protecting the Analog Input

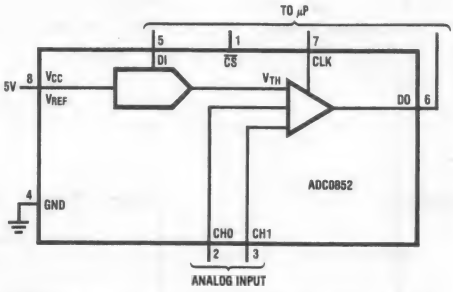

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FIGURE 13. One Component Window Comparator

Requires no additional parts. Window comparisons can be accomplished by inputting the upper and lower window limits into DI on successive comparisons and observing the two outputs:

Two high outputs → input > window

Two low outputs → input < window

One low and one high → input is within window

Typical Applications (Continued)

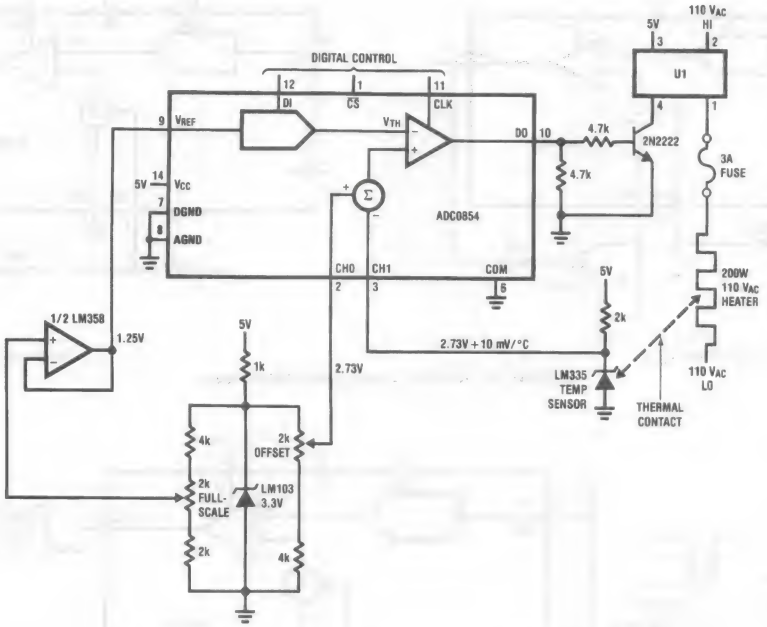

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FIGURE 14. Serial Input Temperature Controller

Note 1: ADC0854 does not require constant service from computer. Self controlled after one write to DI if $\overline{CS}$ remains low.

Note 2: U_1: Solid State Relay, Potter Brumfield #EOM1DB22

Note 3: Set Temp via. DI. Range: 0 to 125°C

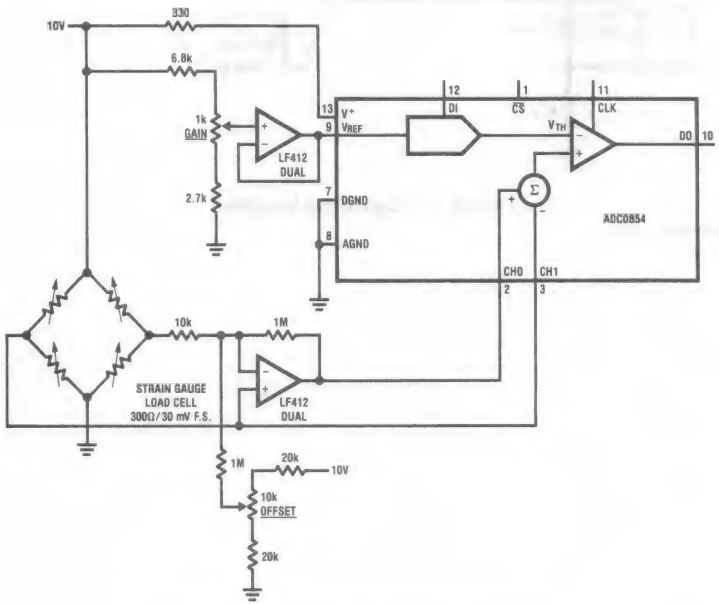

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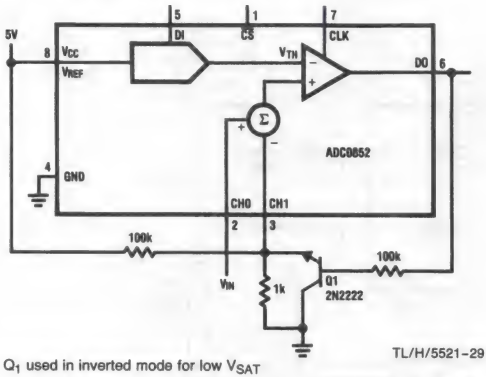
FIGURE 15. Load Cell Limit Comparator

- Differential Input elliminates need for instrumentation amplifier
- A total of 4 load cells can be monitored by ADC0854

Typical Applications (Continued)



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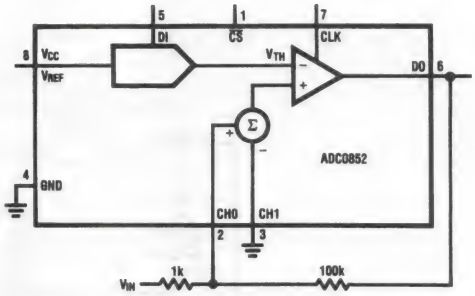

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* Q_1 used in inverted mode for low V_{SAT}

Hysteresis band = 50 mV

FIGURE 16. Adding Comparator Hysteresis

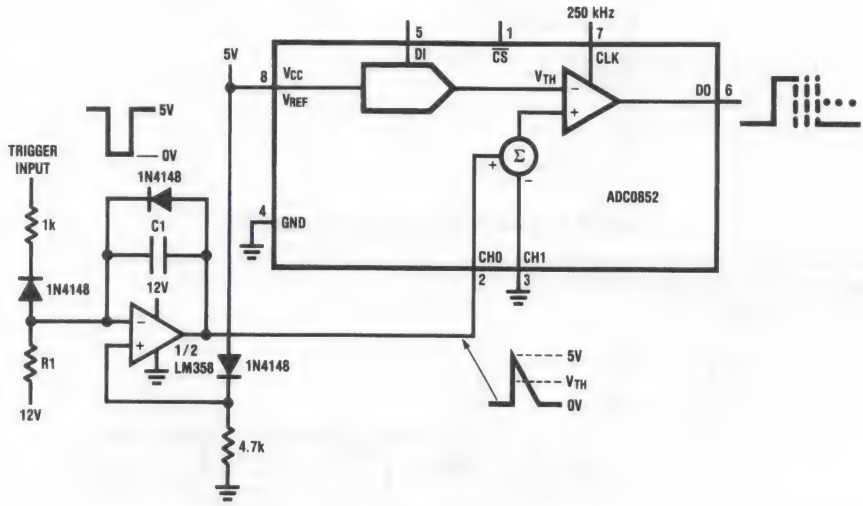

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FIGURE 17. Pulse-Width Modulator

• Range of pulse-widths controlled via R_1, C_1

Typical Applications (Continued)

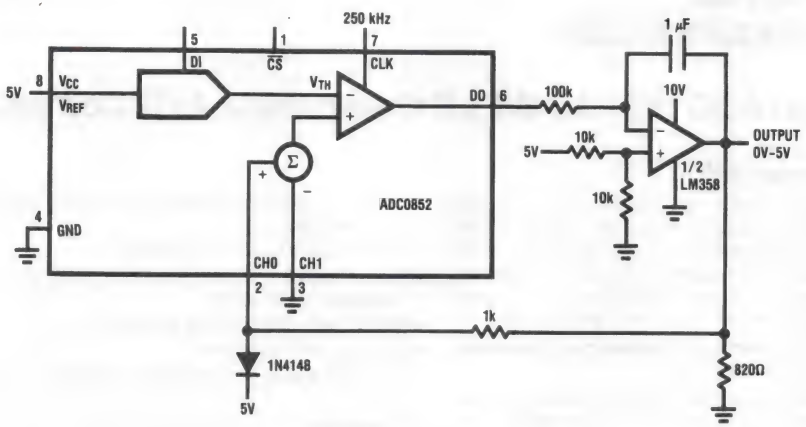

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FIGURE 18. Serial Input 8-Bit DAC

Ordering Information

Part Number	Analog Input Channels	Total Unadjusted Error	Package	Temperature Range
ADC0852BJ	2	±½	J08A	−55°C to +125°C
ADC0852BCJ				−40°C to +85°C
ADC0852BCN			N08E	0°C to 70°C
ADC0852CCJ		±1	J08A	−40°C to +85°C
ADC0852CCN			N08E	0°C to 70°C
ADC0854BJ	4	±½	J14A	−55°C to +125°C
ADC0854BCJ				−40°C to +85°C
ADC0854BCN			N14A	0°C to 70°C
ADC0854CCJ		±1	J14A	−40°C to +85°C
ADC0854CCN			N14A	0°C to 70°C

National Semiconductor

ADC1001/ADC1021 10-Bit μP Compatible A/D Converters

General Description

The ADC1001 and ADC1021 are CMOS, 10-bit successive approximation A/D converters. The 20-pin ADC1001 is pin compatible with the ADC0801 8-bit A/D family. The 10-bit data word is read in two 8-bit bytes, formatted left justified and high byte first. The six least significant bits of the second byte are set to zero, as is proper for a 16-bit word.

The 24-pin ADC1021 outputs 10 bits parallel and is intended for interface to a 16-bit data bus.

Differential inputs provide low frequency input common mode rejection and allow offsetting the analog range of the converter. In addition, the reference input can be adjusted enabling the conversion of reduced analog ranges with 10-bit resolution.

Features

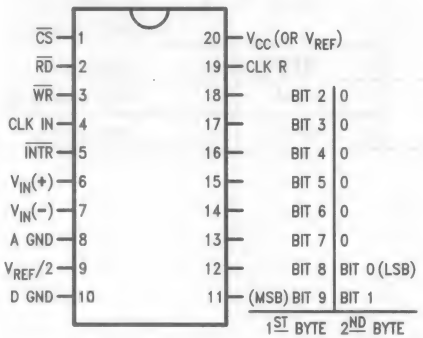
- ADC1001 is pin compatible with ADC0801 series 8-bit A/D converters
- Compatible with NSC800 and 8080 μP derivatives—no interfacing logic needed
- Easily interfaced to 6800 μP derivatives with minimal external logic
- Differential analog voltage inputs
- Logic inputs and outputs meet both MOS and TTL voltage level specifications
- Works with 2.5V (LM336) voltage reference
- On-chip clock generator
- 0V to 5V analog input voltage range with single 5V supply
- Operates ratiometrically or with 5 V_{DC}, 2.5 V_{DC}, or analog span adjusted voltage reference
- 0.3" standard width 20-pin DIP package or 24 pins with 10-bit parallel output

Key Specifications

■ Resolution	10 bits
■ Linearity error	±1 LSB
■ Conversion time	200μS

Connection Diagrams

ADC1001 (for an 8-bit data bus)
Dual-In-Line Package



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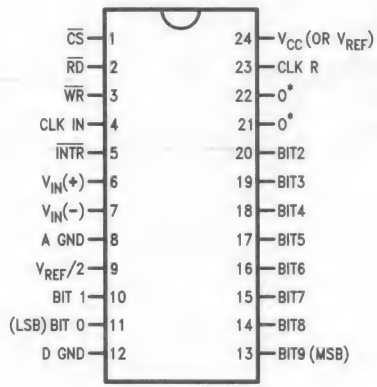
Top View

ADC1021 (for all 10-bit outputs in parallel)
Dual-In-Line Package



TL/H/5675–12

Top View

*TRI-STATE output buffers which output 0 during $\overline{RD}$.

Ordering Information

Temperature Range	0°C to +70°C		−40°C to +85°C	
Order Number	ADC1001CCJ-1	ADC1021CCJ-1	ADC1001CCJ	ADC1021CCJ
Package Outline	J20A	J24A	J20A	J24A

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC}) (Note 3)	6.5V
Logic Control Inputs	−0.3V to +18V
Voltage at Other Inputs and Outputs	−0.3V to (V_{CC}+0.3V)
Storage Temperature Range	−65°C to +150°C
Package Dissipation at T_A=25°C	875 mW
Lead Temp. (Soldering, 10 seconds)	300°C
ESD Susceptibility (Note 10)	800V

Operating Conditions (Notes 1 & 2)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC1001CCJ	−40°C≤T_A≤+85°C
ADC1021CCJ	
ADC1001CCJ-1	0°C≤T_A≤+70°C
ADC1021CCJ-1	
Range of V_{CC}	4.5 V_{DC} to 6.3 V_{DC}

Converter Characteristics

Converter Specifications: V_{CC}=5 V_{DC}, $V_{REF}/2$=2.500 V_{DC}, $T_{MIN} \le T_A \le T_{MAX}$ and f_{CLK}=410 kHz unless otherwise specified.

Parameter	Conditions	Min	Typ	Max	Units
ADC1001C, ADC1021C:					
Linearity Error				±1	LSB
Zero Error				±2	LSB
Full-Scale Error				±2	LSB
Total Ladder Resistance (Note 9)	Input Resistance at Pin 9	2.2	4.8		KΩ
Analog Input Voltage Range	(Note 4) V(+) or V(−)	GND−0.05		V_{CC}+0.05	V_{DC}
DC Common-Mode Error	Over Analog Input Voltage Range		±⅛		LSB
Power Supply Sensitivity	V_{CC}=5 V_{DC}±5% Over Allowed V_{IN}(+) and V_{IN}(−) Voltage Range (Note 4)		±⅛		LSB

AC Electrical Characteristics

Timing Specifications: V_{CC}=5 V_{DC} and T_A=25°C unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
T_c	Conversion Time	(Note 5) f_{CLK}=410 kHz	80 196		90 219	$1/f_{CLK}$ μS
f_{CLK}	Clock Frequency	(Note 8)	100		1260	kHz
	Clock Duty Cycle		40		60	%
CR	Conversion Rate In Free-Running Mode	$\overline{INTR}$ tied to $\overline{WR}$ with $\overline{CS}$=0 V_{DC}, f_{CLK}=410 kHz			4600	conv/s
$t_{W(\overline{WR})L}$	Width of $\overline{WR}$ Input (Start Pulse Width)	$\overline{CS}$=0 V_{DC} (Note 6)	150			ns
t_{ACC}	Access Time (Delay from Falling Edge of $\overline{RD}$ to Output Data Valid)	C_L=100 pF		170	300	ns
t_{1H}, t_{0H}	TRI-STATE® Control (Delay from Rising Edge of $\overline{RD}$ to Hi-Z State)	C_L=10 pF, R_L=10k (See TRI-STATE Test Circuits)		125	200	ns
t_{WI}, t_{RI}	Delay from Falling Edge of $\overline{WR}$ or $\overline{RD}$ to Reset of $\overline{INTR}$			300	450	ns
t_{1rs}	$\overline{INTR}$ to 1st Read Set-Up Time		550	400		ns
C_{IN}	Input Capacitance of Logic Control Inputs			5	7.5	pF
C_{OUT}	TRI-STATE Output Capacitance (Data Buffers)			5	7.5	pF

3

DC Electrical Characteristics

The following specifications apply for $V_{CC}=5\ V_{DC}$ and $T_{MIN}\leq T_A\leq T_{MAX}$, unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
CONTROL INPUTS [Note: CLK IN is the input of a Schmitt trigger circuit and is therefore specified separately]						
V_{IN} (1)	Logical "1" Input Voltage (Except CLK IN)	$V_{CC}=5.25\ V_{DC}$	2.0		15	V_{DC}
V_{IN} (0)	Logical "0" Input Voltage (Except CLK IN)	$V_{CC}=4.75\ V_{DC}$			0.8	V_{DC}
I_{IN} (1)	Logical "1" Input Current (All Inputs)	$V_{IN}=5\ V_{DC}$		0.005	1	μA_{DC}
I_{IN} (0)	Logical "0" input Current (All Inputs)	$V_{IN}=0\ V_{DC}$	−1	−0.005		μA_{DC}
CLOCK IN						
V_T+	CLK IN Positive Going Threshold Voltage		2.7	3.1	3.5	V_{DC}
V_T-	CLK IN Negative Going Threshold Voltage		1.5	1.8	2.1	V_{DC}
V_H	CLK IN Hysteresis $(V_T+)-(V_T-)$		0.6	1.3	2.0	V_{DC}
OUTPUTS AND $\overline{INTR}$						
$V_{OUT}(0)$	Logical "0" Output Voltage	$I_{OUT}=1.6$ mA, $V_{CC}=4.75\ V_{DC}$			0.4	V_{DC}
$V_{OUT}(1)$	Logical "1" Output Voltage	$I_O=-360\ \mu A$, $V_{CC}=4.75\ V_{DC}$ $I_O=-10\ \mu A$, $V_{CC}=4.75\ V_{DC}$	2.4 4.5			V_{DC} V_{DC}
I_{OUT}	TRI-STATE Disabled Output Leakage (All Data Buffers)	$V_{OUT}=0.4\ V_{DC}$ $V_{OUT}=5\ V_{DC}$		0.1 0.1	−100 3	μA_{DC} μA_{DC}
I_{SOURCE}		V_{OUT} Short to GND, $T_A=25°C$	4.5	6		mA_{DC}
I_{SINK}		V_{OUT} Short to V_{CC}, $T_A=25°C$	9.0	16		mA_{DC}
POWER SUPPLY						
I_{CC}	Supply Current (Includes Ladder Current)	$f_{CLK}=410$ kHz, $V_{REF}/2=$ NC, $T_A=25°C$ and $\overline{CS}=1$		2.5	5.0	mA

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to GND, unless otherwise specified. The separate A GND point should always be wired to the D GND.

Note 3: A zener diode exists, internally, from V_{CC} to GND and has a typical breakdown voltage of 7 V_{DC}.

Note 4: For $V_{IN}(-)\geq V_{IN}(+)$ the digital output code will be all zeros. Two on-chip diodes are tied to each analog input (see Block Diagram) which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct—especially at elevated temperatures, and cause errors for analog inputs near fullscale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

Note 5: With an asynchronous start pulse, up to 8 clock periods may be required before the internal clock phases are proper to start the conversion process. The start request is internally latched, see *Figure 1*.

Note 6: The $\overline{CS}$ input is assumed to bracket the $\overline{WR}$ strobe input and therefore timing is dependent on the $\overline{WR}$ pulse width. An arbitrarily wide pulse width will hold the converter in a reset mode and the start of conversion is initiated by the low to high transition of the $\overline{WR}$ pulse (see Timing Diagrams).

Note 7: All typical values are for $T_A=25°C$.

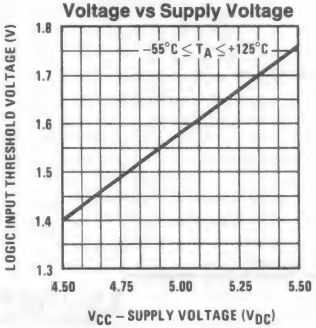
Note 8: Accuracy is guaranteed at $f_{CLK}=410$ kHz. At higher clock frequencies accuracy can degrade.

Note 9: The $V_{REF/2}$ pin is the center point of a two resistor divider (each resistor is 2.4kΩ) connected from V_{CC} to ground. Total ladder input resistance is the sum of these two equal resistors.

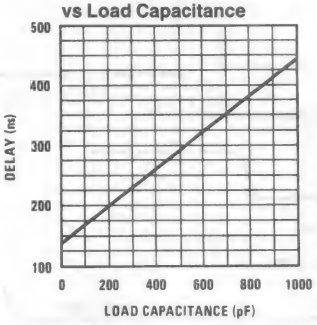
Note 10: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Typical Performance Characteristics

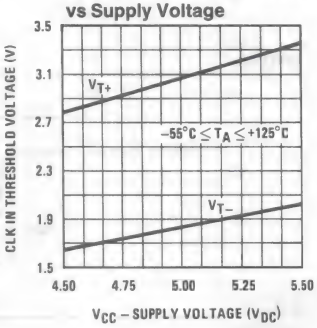
Logic Input Threshold Voltage vs Supply Voltage



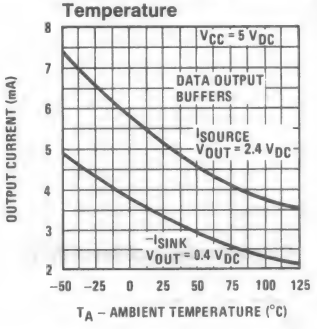
Delay From Falling Edge of $\overline{RD}$ to Output Data Valid vs Load Capacitance



CLK IN Schmitt Trip Levels vs Supply Voltage

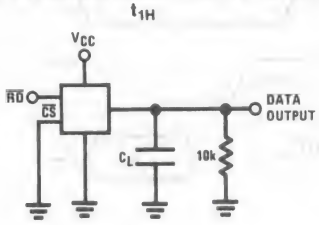


Output Current vs Temperature

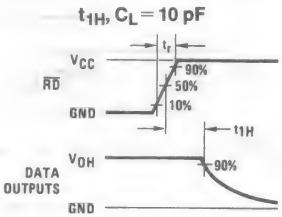


TL/H/5675–2

TRI-STATE Test Circuits and Waveforms

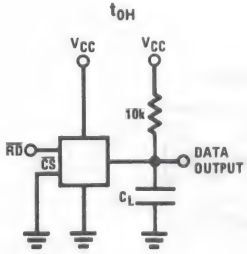


TL/H/5675–3

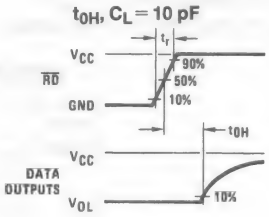


TL/H/5675–4

$t_r = 20$ ns



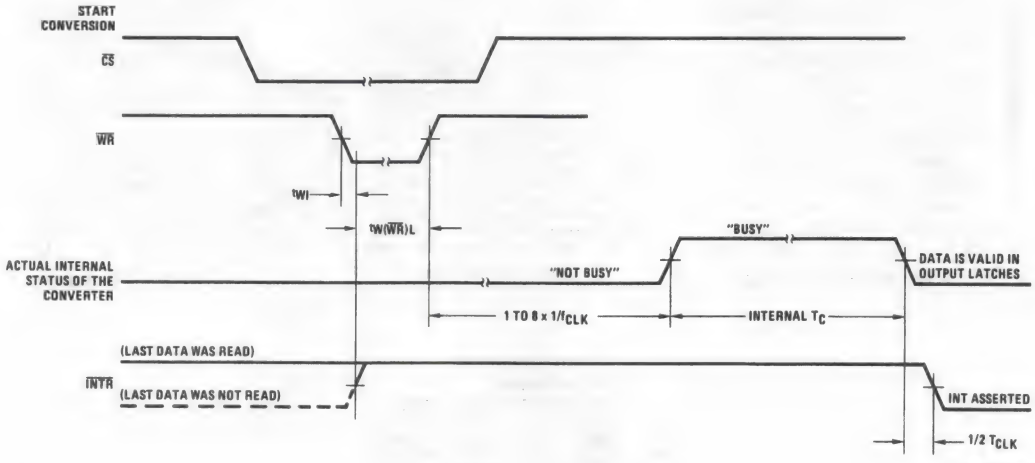
TL/H/5675–5



TL/H/5675–6

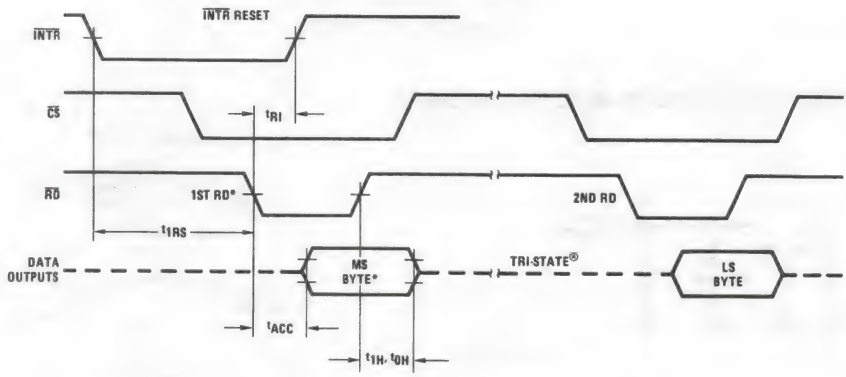
$t_r = 20$ ns

Timing Diagrams



TL/H/5675–7

Output Enable and Reset $\overline{INTR}$



TL/H/5675–8

*The 24-pin ADC1021 outputs all 10 bits on each RD.

Note: All timing is measured from the 50% voltage points.

BYTE SEQUENCING FOR THE 20-PIN ADC1001

Byte Order	8-Bit Data Bus Connection							
	DB7	DB6	DB5	DB4	DB3	DB2	DB1	DB0
1st	MSB Bit 9	Bit 8	Bit 7	Bit 6	Bit 5	Bit 4	Bit 3	Bit 2
2nd	Bit 1	LSB Bit 0	0	0	0	0	0	0

Functional Description

The ADC1001, ADC1021 use an advanced potentiometric resistive ladder network. The analog inputs, as well as the taps of this ladder network, are switched into a weighted capacitor array. The output of this capacitor array is the input to a sampled data comparator. This comparator allows the successive approximation logic to match the analog difference input voltage $[V_{IN}(+)-V_{IN}(-)]$ to taps on the R network. The most significant bit is tested first and after 10 comparisons (80 clock cycles) a digital 10-bit binary code (all "1"s = full-scale) is transferred to an output latch and then an interrupt is asserted ($\overline{INTR}$ makes a high-to-low transition). The device may be operated in the free-running mode by connecting $\overline{INTR}$ to the $\overline{WR}$ inut with $\overline{CS}=0$. To ensure start-up under all possible conditions, an external $\overline{WR}$ pulse is required during the first power-up cycle. A conversion in process can be interrupted by issuing a second start command.

On the high-to-low transition of the $\overline{WR}$ input the internal SAR latches and the shift register stages are reset. As long as the $\overline{CS}$ input and $\overline{WR}$ input remain low, the A/D will remain in a reset state. *Conversion will start from 1 to 8 clock periods after at least one of these inputs makes a low-to-high transition.*

A functional diagram of the A/D converter is shown in *Figure 1*. All of the inputs and outputs are shown and the major logic control paths are drawn in heavier weight lines.

The conversion is initialized by taking $\overline{CS}$ and $\overline{WR}$ simultaneously low. This sets the start flip-flop (F/F) and the resulting "1" level resets the 8-bit shift register, resets the Interrupt (INTR) F/F and inputs a "1" to the D flop, F/F1, which is at the input end of the 10-bit shift register. Internal clock signals then transfer this "1" to the Q output of F/F1. The AND gate, G1, combines this "1" output with a clock signal to provide a reset signal to the start F/F. If the set signal is no longer present (either $\overline{WR}$ or $\overline{CS}$ is a "1") the start F/F is reset and the 10-bit shift register then can have the "1" clocked in, which allows the conversion process to continue. If the set signal were to still be present, this reset pulse would have no effect and the 10-bit shift register would continue to be held in the reset mode. This logic therefore allows for wide $\overline{CS}$ and $\overline{WR}$ signals and the converter will start after at least one of these signals returns high and the internal clocks again provide a reset signal for the start F/F.

After the "1" is clocked through the 10-bit shift register (which completes the SAR search) it causes the new digital word to transfer to the TRI-STATE output latches. When this XFER signal makes a high-to-low transition the one shot fires, setting the INTR F/F. An inverting buffer then supplies the $\overline{INTR}$ output signal.

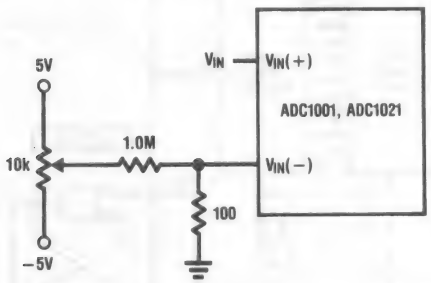
Note that this $\overline{SET}$ control of the INTR F/F remains low for aproximately 400 ns. If the data output is continuously enabled ($\overline{CS}$ and $\overline{RD}$ both held low), the $\overline{INTR}$ output will still signal the end of the conversion (by a high-to-low transition), because the $\overline{SET}$ input can control the Q output of the INTR F/F even though the RESET input is constantly at a "1" level. This $\overline{INTR}$ output will therefore stay low for the duration of the $\overline{SET}$ signal.

When data is to be read, the combination of both $\overline{CS}$ and $\overline{RD}$ being low will cause the INTR F/F to be reset and the TRI-STATE output latches will be enabled.

Zero and Full-Scale Adjustment

Zero error can be adjusted as shown in *Figure 2*. $V_{IN}(+)$ is forced to +2.5 mV (+½ LSB) and the potentiometer is adjusted until the digital output code changes from 00 0000 0000 to 00 0000 0001.

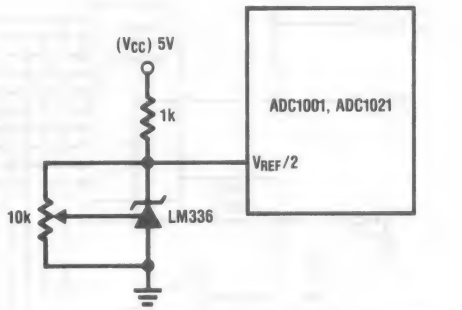
Full-scale is adjusted as shown in *Figure 3*, with the $V_{REF}/2$ input. With V_{IN} (+) forced to the desired full-scale voltage less 1½ LSBs ($V_{FS}-1½$ LSBs), $V_{REF}/2$ is adjusted until the digital output code changes from 11 1111 1110 to 11 1111 1111.



TL/H/5675–9

NOTE: $V_{IN}(-)$ should be biased so that $V_{IN}(-) \geq -0.05V$ when potentiometer wiper is set at most negative voltage position.

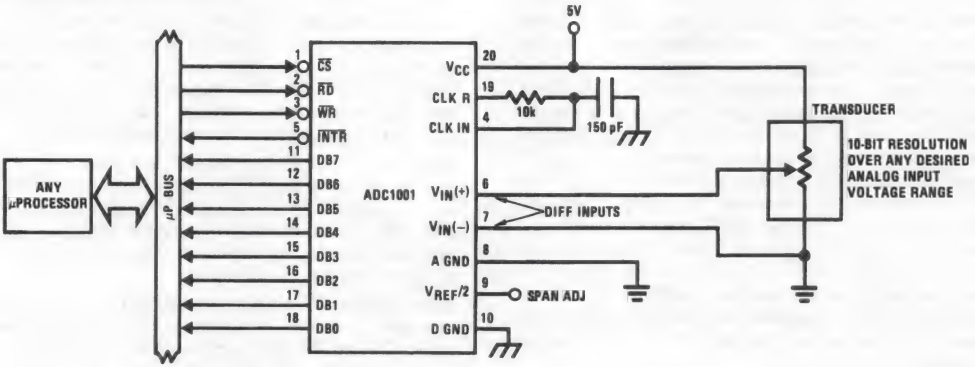
FIGURE 2. Zero Adjust Circuit



TL/H/5675–10

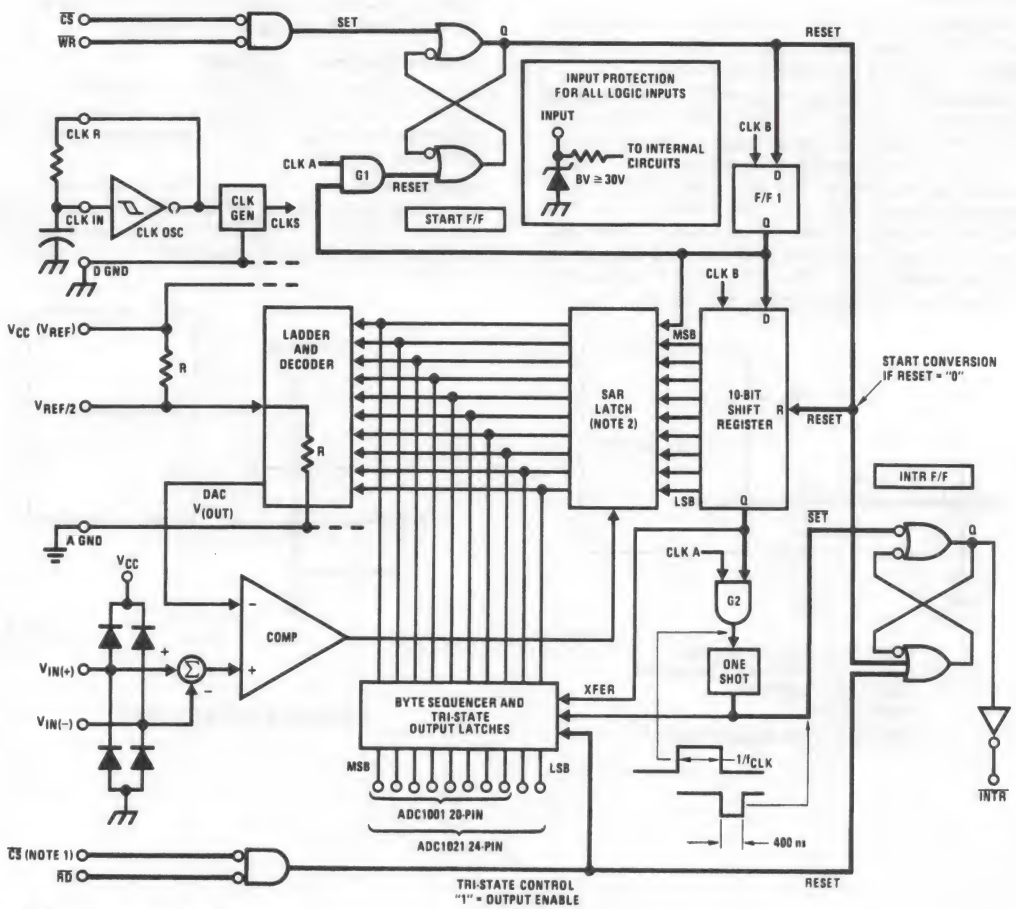
FIGURE 3. Full-Scale Adjust

Typical Application



TL/H/5675–1

Block Diagram



Note 1: $\overline{CS}$ shown twice for clarity.

Note 2: SAR = Successive Approximation Register.

TL/H/5675–13

FIGURE 1

National Semiconductor

ADC1005/ADC1025 10-Bit μP Compatible A/D Converters

General Description

The ADC1005 and ADC1025 are CMOS 10-bit successive approximation A/D converters. The 20-pin ADC1005 outputs 10-bit data in a two-byte format for interface with 8-bit microprocessors.

The 24-pin ADC1025 outputs 10 bits in parallel and is intended for 16-bit data buses or stand-alone applications.

Both A-to-Ds have differential inputs to permit rejection of common-mode signals, allow the analog input range to be offset, and also to permit the conversion of signals not referred to ground. In addition, the reference voltage can be adjusted, allowing smaller voltage spans to be measured with 10-bit resolution.

Features

- Easy interface to all microprocessors
- Differential analog voltage inputs
- Operates ratiometrically or with 5 V_{DC} voltage reference or analog span adjusted voltage reference
- 0V to 5V analog input voltage range with single 5V supply
- On-chip clock generator
- TLL/MOS input/output compatible
- 0.3" standard width 20-pin DIP or 24-pin DIP with 10-bit parallel output
- Available in 20-pin or 28-pin molded chip carrier package

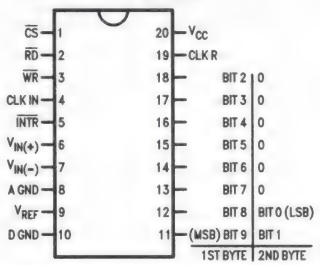
Key Specifications

■ Resolution	10 bits
■ Linearity Error	±½ LSB and ±1 LSB
■ Conversion Time	50 μs

Connection Diagrams

ADC1005 (for an 8-bit data bus)

Dual-In-Line Package

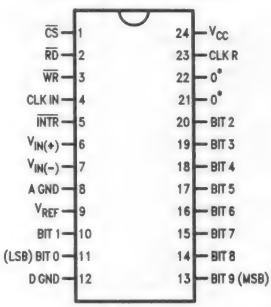


TL/H/5261–1

Top View

ADC1025 (10-bit parallel outputs)

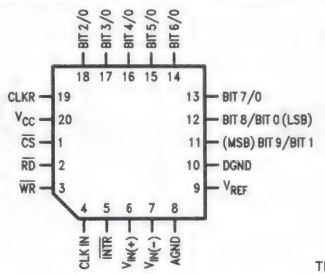
Dual-In-Line Package



TL/H/5261–2

Top View

ADC1005 Molded Chip Carrier Package



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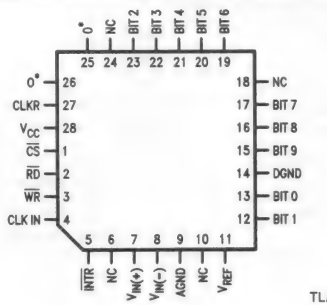
Top View

ADC1025 Molded Chip Carrier Package



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Top View

*TRI-STATE® output buffers which output 0 during $\overline{RD}$

See Ordering Information

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	6.5V
Logic Control Inputs	−0.3V to +15V
Voltage at Other Inputs and Outputs	−0.3V to V_{CC} +0.3V
Input Current Per Pin	±5 mA
Input Current Per Package	±20 mA
Storage Temperature Range	−65°C to +150°C
Package Dissipation at T_A = 25°C	875 mW
Lead Temperature	
(Soldering, 10 seconds)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Surface Mount Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C
ESD Susceptibility (Note 8)	800V

Operating Ratings (Notes 1 & 2)

Supply Voltage (V_{CC})	4.5V to 6.0V
Temperature Range	$T_{MN} \le T_A \le T_{MAX}$
ADC1005BJ, ADC1005CJ	−55°C ≤ T_A ≤ +125°C
ADC1025BJ, ADC1025CJ	
ADC1005BCJ, ADC1005CCJ	−40°C ≤ T_A ≤ +85°C
ADC1025BCJ, ADC1025CCJ	
ADC1005BCJ-1, ADC1005CCJ-1	0°C ≤ T_A ≤ 70°C
ADC1025BCJ-1, ADC1025CCJ-1	
ADC1005BCN, ADC1005CCN	
ADC1025BCN, ADC1025CCN	
ADC1005BCV, ADC1005CCV	
ADC1025BCV, ADC1025CCV	

Electrical Characteristics

The following specifications apply for V_{CC} = 5V, V_{REF} = 5V, f_{CLK} = 1.8 MHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** All other limits T_A = T_j = 25°C.

Parameter	Conditions	ADC10X5BJ, ADC10X5BCJ ADC10X5CJ, ADC10X5CCJ			ADC10X5BCJ-1, ADC10X5CCJ-1 ADC10X5BCN, ADC10X5CCN ADC10X5BCV, ADC10X5CCV			Limit Units
		Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	
Converter Characteristics								
Linearity Error (Note 3)								
ADC10X5BJ, ADC10X5BCJ			**±0.5**					LSB
ADC10X5BCJ-1, BCN, BCV						±0.5	**±0.5**	LSB
ADC10X5CJ, ADC10X5CCJ			**±1**					LSB
ADC10X5CCJ-1, CCN, CCV						±1	**±1**	LSB
Zero Error								
ADC10X5BJ, ADC10X5BCJ			**±0.5**					LSB
ADC10X5BCJ-1, BCN, BCV						±0.5	**±0.5**	LSB
ADC10X5CJ, ADC10X5CCJ			**±1**					LSB
ADC10X5CCJ-1, CCN, CCV						±1	**±1**	LSB
Fullscale Error								
ADC10X5BJ, ADC10X5BCJ			**±0.5**					LSB
ADC10X5BCJ-1, BCN, BCV						±0.5	**±0.5**	LSB
ADC10X5CJ, ADC10X5CCJ			**±1**					LSB
ADC10X5CCJ-1, CCN, CCV						±1	**±1**	LSB
Reference Input Resistance MIN		4.8	**2.2**		4.8	2.4	**2.2**	kΩ
Reference Input Resistance MAX		4.8	**8.3**		4.8	7.6	**8.3**	kΩ
Common-Mode Input (Note 4) MIN	$V_{IN}(+)$ or $V_{IN}(-)$		**V_{CC} + 0.05**			V_{CC} + 0.05	**V_{CC} + 0.05**	V
Common-Mode Input (Note 4) MAX	$V_{IN}(+)$ or $V_{IN}(-)$		**GND − 0.05**			GND − 0.05	**GND − 0.05**	V
DC Common-Mode Error	Over Common-Mode Input Range	±⅛	±¼		±⅛	±¼	±¼	LSB
Power Supply Sensitivity	V_{CC} = 5 V_{DC} ±5% V_{REF} = 4.75V	±⅛	±¼		±⅛	±¼	±¼	LSB

Electrical Characteristics (Continued) The following specifications apply for V_{CC} = 5V, V_{REF} = 5V, f_{CLK} = 1.8 MHz unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** All other limits $T_A = T_j$ = 25°C.

Parameter	Conditions	ADC10X5BJ, ADC10X5BCJ ADC10X5CJ, ADC10X5CCJ			ADC10X5BCJ-1, ADC10X5CCJ-1 ADC10X5BCN, ADC10X5CCN ADC10X5BCV, ADC10X5CCV			Limit Units
		Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	
DC Characteristics								
$V_{IN(1)}$ Logical "1" Input Voltage MIN	V_{CC} = 5.25V (except CLK_{IN})		**2.0**			2.0	**2.0**	V
$V_{IN(0)}$, Logical "0" Input Voltage MAX	V_{CC} = 4.75V (Except CLK_{IN})		**0.8**			0.8	**0.8**	V
I_{IN}, Logical "1" Input Current MAX	V_{IN} = 5.0V	0.005	**1**		0.005	1	**1**	μA
I_{IN}, Logical "0" Input Current MAX	V_{IN} = 0V	−0.005	**−1**		−0.005	−1	**−1**	μA
$V_{T+(MIN)}$, Minimum CLK_{IN} Positive going Threshold Voltage		3.1	**2.7**		3.1	2.7	**2.7**	V
$V_{T(MAX)}$, Maximum CLK_{IN} Positive going Threshold Voltage		3.1	**3.5**		3.1	3.5	**3.5**	V
$V_{T-(MIN)}$, Minimum CLK_{IN} Negative going Threshold Voltage		1.8	**1.5**		1.8	1.5	**1.5**	V
$V_{T-(MAX)}$, Maximum CLK_{IN} Negative going Threshold Voltage		1.8	**2.1**		1.8	2.1	**2.1**	V
$V_{H(MIN)}$, Minimum CLK_{IN} Hysteresis (V_{T+}-V_{T-})		1.3	**0.6**		1.3	0.6	**0.6**	V
$V_{H(MAX)}$, Maximum CLK_{IN} Hysteresis (V_{T+}-V_{T-})		1.3	**2.0**		1.3	2.0	**2.0**	V
$V_{OUT(1)}$, Logical "1" Output Voltage MIN	V_{CC} = 4.75V I_{OUT} = −360 μA I_{OUT} = −10 μA		**2.4** **4.5**			2.8 4.6	**2.4** **4.5**	V V
$V_{OUT(0)}$, Logical "0" Output Voltage MAX	V_{CC} = 4.75V I_{OUT} = 1.6 mA		**0.4**			0.34	**0.4**	V
I_{OUT}, TRI-STATE Output Current MAX	V_{OUT} = 0V V_{OUT} = 5V	−0.01 0.01	**−3** **3**		−0.01 0.01	−0.3 0.3	**−3** **3**	μA μA
I_{SOURCE}, Output Source Current MIN	V_{OUT} = 0V	−14	**−6.5**		−14	−7.5	**−6.5**	mA
I_{SINK}, Output Sink Current MIN	V_{OUT} = 5V	16	**8.0**		16	9.0	**8.0**	mA
I_{CC}, Supply Current MAX	f_{CLK} = 1.8 MHz $\overline{CS}$ = "1"	1.5	**3**		1.5	2.5	**3**	mA

AC Electrical Characteristics The following specifications apply for V_{CC} = 5V, V_{REF} = 5V, $t_r = t_f$ = 20 ns unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** All other limits $T_A = T_j$ = 25°C.

Parameter	Conditions	Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	Limit Units
f_{CLK}, Clock Frequency MIN			0.2	0.2	MHz
MAX			2.6	2.6	MHz
Clock Duty Cycle MIN			40	40	%
MAX			60	60	%

AC Electrical Characteristics

The following specifications apply for V_{CC} = 5V, V_{REF} = 5V, $t_r = t_f$ = 20 ns unless otherwise specified. **Boldface limits apply from T_{MIN} to T_{MAX};** All other limits $T_A = T_j$ = 25°C. (Continued)

Parameter		Conditions	Typ (Note 5)	Tested Limit (Note 6)	Design Limit (Note 7)	Limit Units
t_C, Conversion Time	MIN			80	80	$1/f_{CLK}$
	MAX			90	90	$1/f_{CLK}$
	MIN	f_{CLK} = 1.8 MHz		45	45	μs
	MAX	f_{CLK} = 1.8 MHz		50	50	μs
$t_{W(\overline{WR})L}$, Minimum $\overline{WR}$ Pulse Width		$\overline{CS}$ = 0	100	150	150	ns
t_{ACC}, Access Time (Delay from falling edge of $\overline{RD}$ to Output Data Valid)		$\overline{CS}$ = 0 C_L = 100 pF, R_L = 2k	170	300	300	ns
t_{1H}, t_{0H}, TRI-STATE Control (Delay from Rising Edge of $\overline{RD}$ to Hi-Z State)		R_L = 10k, C_L = 10 pF R_L = 2k, C_L = 100 pF	125 145	 230	200 230	ns ns
t_{WI}, t_{RI}, Delay from Falling Edge of $\overline{WR}$ or $\overline{RD}$ to Reset of $\overline{INTR}$			300	450	450	ns
t_{IRS}, $\overline{INTR}$ to 1st Read Set-up Time			400	550	550	ns
C_{IN}, Capacitance of Logic Inputs			5		7.5	pF
C_{OUT}, Capacitance of Logic Outputs			5		7.5	pF

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to ground.

Note 3: Linearity error is defined as the deviation of the analog value, expressed in LSBs, from the straight line which passes through the end points of the transfer characteristic.

Note 4: For $V_{IN(-)} \geq V_{IN(+)}$ the digital output code will be 00 0000 0000. Two on-chip diodes are tied to each analog input which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct, especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50 mV forward bias of either diode. This means that as long as the analog V_{IN} does not exceed the supply voltage by more than 50 mV, the output code will be correct. To achieve an absolute 0 V_{DC} to 5 V_{DC} input voltage range will therefore require a minimum supply voltage of 4.950 V_{DC} over temperature variations, initial tolerance and loading.

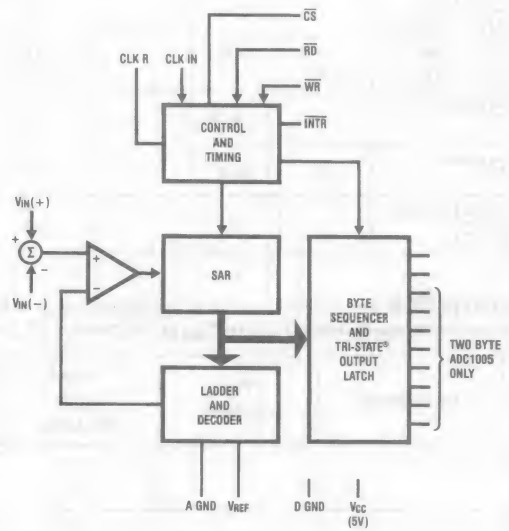
Note 5: Typicals are at 25°C and represent most likely parametric norm.

Note 6: Tested and guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 7: Guaranteed, but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Note 8: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Functional Diagram



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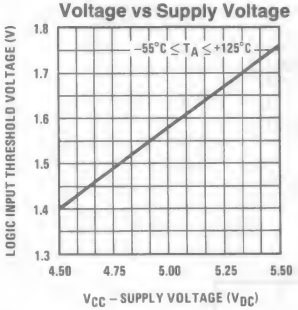
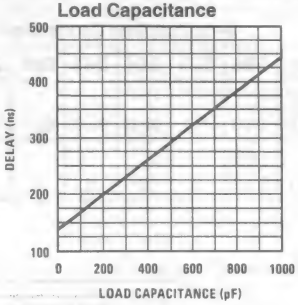
Typical Performance Characteristics

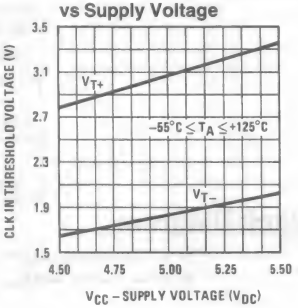
Logic Input Threshold Voltage vs Supply Voltage



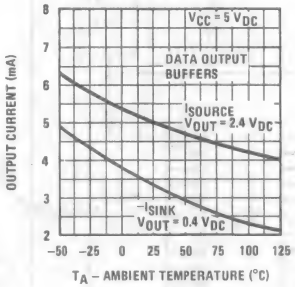
Delay from Falling Edge of $\overline{RD}$ to Output Data Valid vs Load Capacitance



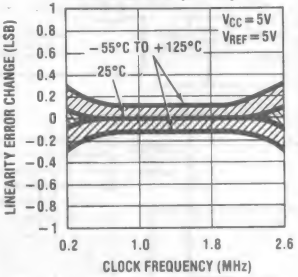
CLK IN Schmitt Trip Levels vs Supply Voltage



Output Current vs Temperature



Typical Linearity Error vs Clock Frequency



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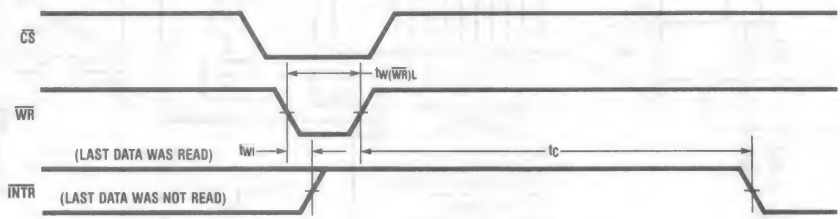
Timing Diagrams

Start Conversion



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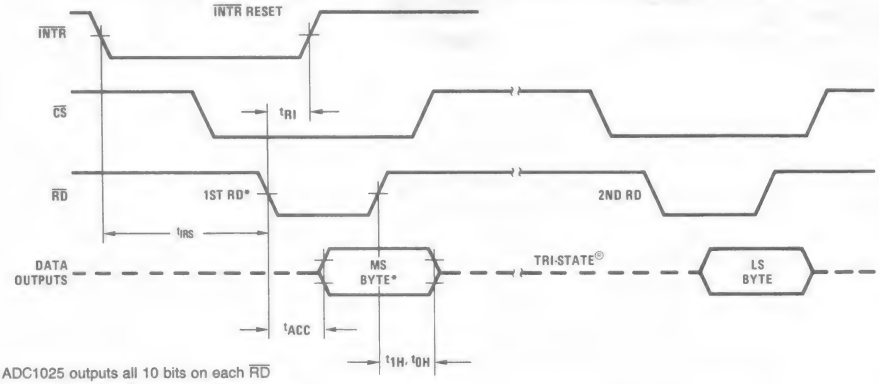
Output Enable and Reset $\overline{INTR}$



*The 24-pin ADC1025 outputs all 10 bits on each $\overline{RD}$

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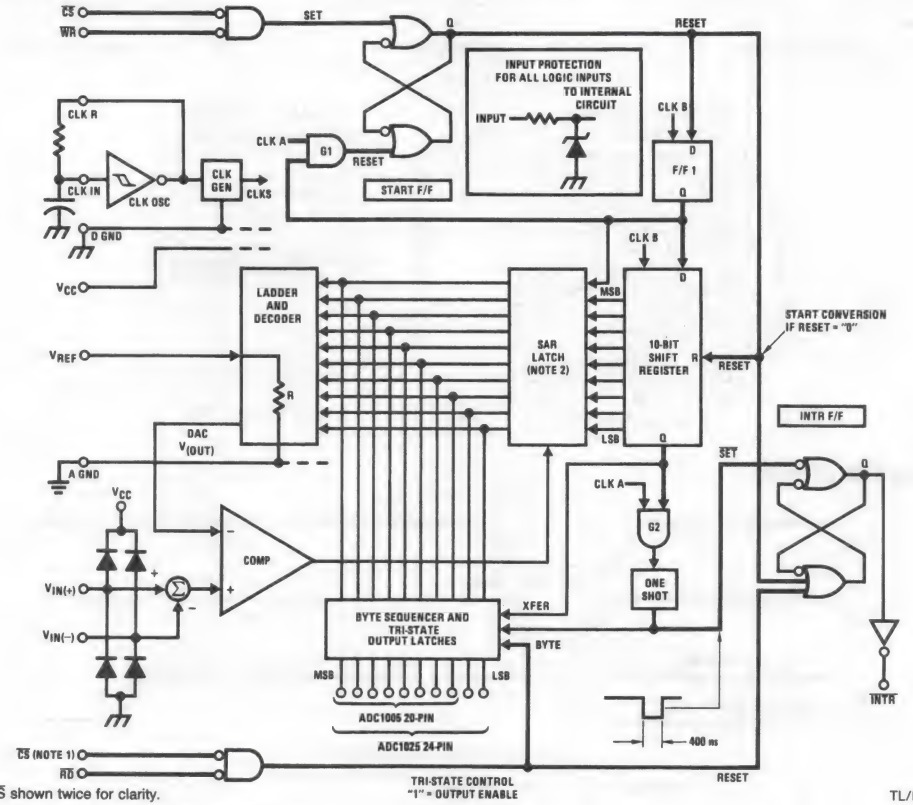
Note: All timing is measured from the 50% voltage points.

Timing Diagrams (Continued)

Byte Sequencing for the 20-Pin ADC1005

Byte Order	8-Bit Data Bus Connection							
	DB7	DB6	DB5	DB4	DB3	DB2	DB1	DB0
1st	MSB Bit 9	Bit 8	Bit 7	Bit 6	Bit 5	Bit 4	Bit 3	Bit 2
2nd	Bit 1	LSB Bit 0	0	0	0	0	0	0

Block Diagram



Note 1: $\overline{CS}$ shown twice for clarity.

Note 2: SAR = Successive Approximation Register.

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FIGURE 1.

Functional Description

1.0 GENERAL OPERATION

A block diagram of the A/D converter is shown in *Figure 1* All of the inputs and outputs are shown and the major logic control paths are drawn in heavier weight lines.

1.1 Converter Operation

The ADC1005, ADC1025 use an advanced potentiometric resistive ladder network. The analog inputs, as well as the taps of this ladder network are switched into a weighted capacitor array. The output of this capacitor array is the input to a sampled data comparator. This comparator allows the successive approximation logic to match the analog input voltage $[V_{IN}(+) - V_{IN}(-)]$ to taps on the R network. The most significant bit is tested first and after 10 comparisons (80 clock cycles) a digital 10-bit binary code (all "1"s = full-scale) is transferred to an output latch.

1.2 Starting a Conversion

The conversion is initialized by taking $\overline{CS}$ and $\overline{WR}$ simultaneously low. This sets the start flip-flop (F/F) and the resulting "1" level resets the 10-bit shift register, resets the interrupt (INTR) F/F and inputs a "1" to the D flop, F/F1, which is at the input end of the 10-bit shift register. Internal clock signals then transfer this "1" to the Q ouput of F/F1. The AND gate, G1, combines this "1" output with a clock signal to provide a reset signal to the start F/F. If the set signal is no longer present (either $\overline{WR}$ or $\overline{CS}$ is a "1") the start F/F is reset and the 10-bit shift register then can have the "1" clocked in, allowing the conversion process to continue. If the set signal were still present, this reset pulse would have no effect and the 10-bit shift register would continue to be held in the reset mode. This logic therefore allows for wide $\overline{CS}$ and $\overline{WR}$ signals. The converter will start after at least one of these signals returns high and the internal clocks again provide a reset signal for the start F/F.

To summarize, on the high-to-low transition of the $\overline{WR}$ input the internal SAR latches and the shift register stages are reset. As long as the $\overline{CS}$ input and $\overline{WR}$ input remain low, the A/D will remain in a reset state. *Conversion will start after at least one of these inputs makes a low-to-high transition.*

1.3 Output Control

After the "1" is clocked through the 10-bit shift register (which completes the SAR search) it causes the new digital word to transfer to the TRI-STATE output latches. When the XFER signal makes a high-to-low transition the one shot fires, setting the INTR F/F. An inverting buffer then supplies the $\overline{INTR}$ output signal.

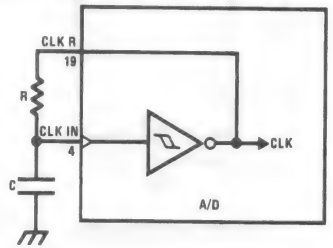
Note that this $\overline{SET}$ control of the INTR F/F remains low for approximately 400 ns. If the data output is continuously enabled ($\overline{CS}$ and $\overline{RD}$ both held low) the $\overline{INTR}$ output will still signal the end of the conversion (by a high-to-low transition). This is because the $\overline{SET}$ input can control the Q output of the INTR F/F even though the RESET input is constantly at a "1" level. This $\overline{INTR}$ output will therefore stay low for the duration of the $\overline{SET}$ signal.

When data is to be read, the combination of both $\overline{CS}$ and $\overline{RD}$ being low will cause the INTR F/F to be reset and the TRI-STATE output latches will be enabled.

1.4 Free-Running and Self-Clocking Modes

For operation in the free-running mode an initializing pulse should be used, following power-up, to ensure circuit operation. In this application, the $\overline{CS}$ input is grounded and the $\overline{WR}$ input is tied to the $\overline{INTR}$ output. This $\overline{WR}$ and $\overline{INTR}$ node should be momentarily forced to logic low following a power-up cycle to ensure start up.

The clock for the A/D can be derived from the CPU clock or an external RC can be added to provide self-clocking. The CLK IN makes use of a Schmitt trigger as shown in *Figure 2.*



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$$f_{CLK} \cong \frac{1}{1.1\,RC}$$

FIGURE 2. Self-Clocking the A/D

2.0 REFERENCE VOLTAGE

The voltage applied to the reference input of these converters defines the voltage span of the analog input (the difference between $V_{IN(MAX)}$ and $V_{IN(MIN)}$) over which the 1024 possible output codes apply. The devices can be used in either ratiometric applications or in systems requiring absolute accuracy. The reference pin must be connected to a voltage source capable of driving the reference input resistance of typically 4.8 kΩ. This pin is the top of a resistor divider string used for the successive approximation conversion.

In a ratiometric system (*Figure 3a*) the analog input voltage is proportional to the voltage used for the A/D reference. This voltage is typically the system power supply, so the V_{REF} pin can be tied to V_{CC}. This technique relaxes the stability requirements of the system references as the analog input and A/D reference move together maintaining the same output code for a given input condition.

For absolute accuracy (*Figure 3b*), where the analog input varies between very specific voltage limits, the reference pin can be biased with a time and temperature stable voltage source. The LM385 and LM336 reference diodes are good low current devices to use with these converters.

The maximum value of the reference is limited to the V_{CC} supply voltage. The minimum value, however, can be small to allow direct conversions of transducer outputs providing less than a 5V output span. Particular care must be taken with regard to noise pickup, circuit layout, and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter (1 LSB equals $V_{REF}/1024$).

Functional Description (Continued)

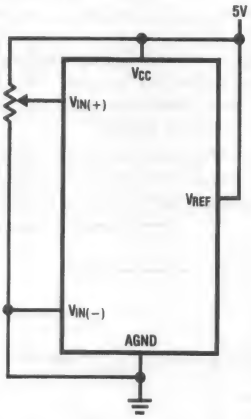

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FIGURE 3a. Ratiometric

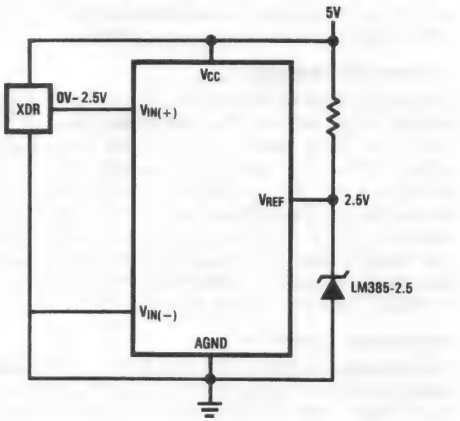

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FIGURE 3b. Absolute with a Reduced Span

3.0 THE ANALOG INPUTS

3.1 Analog Differential Voltage Inputs and Common-Mode Rejection

The differential inputs of these converters reduce the effects of common-mode input noise, which is defined as noise common to both selected "+" and "−" inputs (60 Hz is most typical). The time interval between sampling the "+" input and the "−" input is half of an internal clock period. The change in the common-mode voltage during this short time interval can cause conversion errors. For a sinusoidal common-mode signal, this error is:

$$V_{ERROR(MAX)} = V_{PEAK}\,(2\pi\, f_{CM}) \times \frac{4}{f_{CLK}}$$

where f_{CM} is the frequency of the common-mode signal, V_{PEAK} is its peak voltage value and f_{CLK} is the clock frequency at the CLK IN pin.

For a 60 Hz common-mode signal to generate a 1/4 LSB error (1.2 mV) with the converter running at 1.8 MHz, its peak value would have to be 1.46V. A common-mode signal this large is much greater than that generally found in data aquisition systems.

3.2 Input Current

Due to the sampling nature of the analog inputs, short duration spikes of current enter the "+" input and exit the "−" input at the clock rising edges during the conversion. These currents decay rapidly and do not cause errors as the internal comparator is strobed at the end of a clock period.

3.3 Input Bypass Capacitors

Bypass capacitors at the inputs will average the current spikes noted in 3.2 and cause a DC current to flow through the output resistances of the analog signal sources. This charge pumping action is worse for continuous conversions with the $V_{IN}(+)$ input voltage at full scale. For continuous conversions with a 1.8 MHz clock frequency with the $V_{IN}(+)$ input at 5V, this DC current is at a maximum of approximately 5 μA. Therefore, *bypass capacitors should not be used at the analog inputs or the V_{REF} pin* for high resistance sources (>1 kΩ). If input bypass capacitors are necessary for noise filtering and high source resistance is desirable to minimize capacitor size, the detrimental effects of the voltage drop across this input resistance, which is due to the average value of the input current, can be eliminated with a full-scale adjustment while the given source resistor and input bypass capacitor are both in place. This is possible because the average value of the input current is a linear function of the differential input voltage.

3.4 Input Source Resistance

Large values of source resistance where an input bypass capacitor is not used, *will not cause errors* if the input currents settle out prior to the comparison time. If a low pass filter is required in the system, use a low valued series resistor (≤1 kΩ) for a passive RC section or add an op amp RC active low pass filter. For low source resistance applications (≤0.1 kΩ) a 4700 pF bypass capacitor at the inputs will prevent pickup due to series lead induction of a long wire. A 100Ω series resistor can be used to isolate this capacitor – both the R and the C are placed outside the feedback loop – from the output of an op amp, if used.

3.5 Noise

The leads to the analog inputs (pins 6 and 7) should be kept as short as possible to minimize input noise coupling. Both noise and undesired digital clock coupling to these inputs can cause system errors. The source resistance for these inputs should, in general, be kept below 1 kΩ. Larger values of source resistance can cause undesired system noise pickup. Input bypass capacitors, placed from the analog inputs to ground, can reduce system noise pickup but can create analog scale errors. See section 3.2, 3.3, and 3.4 if input filtering is to be used.

Functional Description (Continued)

4.0 OFFSET AND REFERENCE ADJUSTMENT

4.1 Zero Offset

The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the V(−) input and applying a small magnitude positive voltage to the V(+) input. Zero error is the difference between the actual DC input voltage that is necessary to just cause an output digital code transition from 00 0000 0000 to 00 0000 0001 and the ideal 1/2 LSB value (1/2 LSB = 2.45 mV for V_{REF} = 5.0 V_{DC}).

The zero of the A/D normally does not require adjustment. However, for cases where $V_{IN(MIN)}$ is not ground and in reduced span applications (V_{REF} < 5V), an offset adjustment may be desired. The converter can be made to output an all zero digital code for an arbitrary input by biasing the A/D's $V_{IN}(-)$ input at that voltage. This utilizes the differential input operation of the A/D.

4.2 Full Scale

The full-scale adjustment can be made by applying a differential input voltage that is 1½ LSB down from the desired analog full-scale voltage range and then adjusting the magnitude of the V_{REF} input for a digital output code that is just changing from 11 1111 1110 to 11 1111 1111.

4.3 Adjusting for an Arbitrary Analog Input Voltage Range

If the analog zero voltage of the A/D is shifted away from ground (for example, to accommodate an analog input signal that does not go to ground), this new zero reference should be properly adjusted first. A $V_{IN}(+)$ voltage that equals this desired zero reference plus 1/2 LSB (where the LSB is calculated for the desired analog span, 1 LSB = analog span/1024) is applied to selected "+" input and the zero reference voltage at the corresponding "−" input should then be adjusted to just obtain the 000_{HEX} 001_{HEX} code transition.

The full-scale adjustment should be made [with the proper $V_{IN}(-)$ voltage applied] by forcing a voltage to the $V_{IN}(+)$ input given by:

$$V_{IN}(+)\ \text{FS adj} = V_{MAX} - 1.5\left[\frac{(V_{MAX} - V_{MIN})}{1024}\right]$$

where V_{MAX} = the high end of the analog input range and V_{MIN} = the low end (the offset zero) of the analog range. (Both are ground referenced).

The V_{REF} (or V_{CC}) voltage is then adjusted to provide a code change from $3FF_{HEX}$ to $3FE_{HEX}$. This completes the adjustment procedure.

For an example see the Zero-Shift and Span-Adjust circuit below.

5.0 POWER SUPPLIES

Noise spikes on the V_{CC} supply line can cause conversion errors as the comparator will respond to this noise. A low inductance tantalum filter capacitor should be used close to the converter V_{CC} pin and values of 1 μF or greater are recommended. If an unregulated voltage is available in the system, a separate LM340LAZ-5.0, TO-92, 5V voltage regulator for the converter (and the other analog circuitry) will greatly reduce digital noise on the V_{CC} supply.

A single point analog ground that is separate from the logic ground points should be used. The power supply bypass capacitor and the self-clocking capacitor (if used) should both be returned to the digital ground. Any V_{REF} bypass capacitors, analog input filters capacitors, or input signal shielding should be returned to the analog ground point.

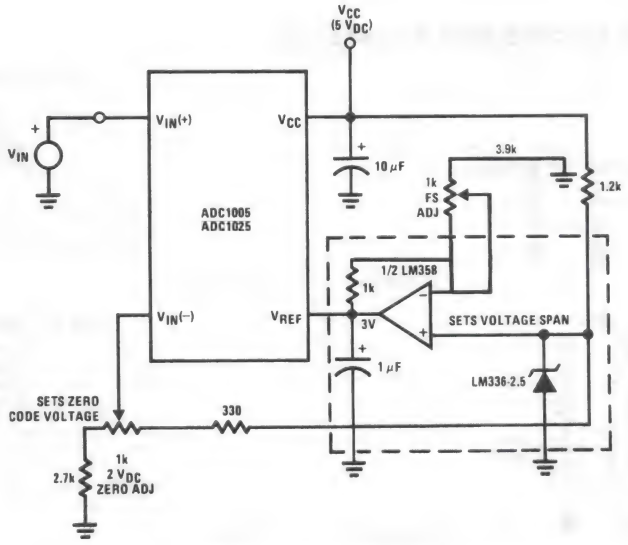

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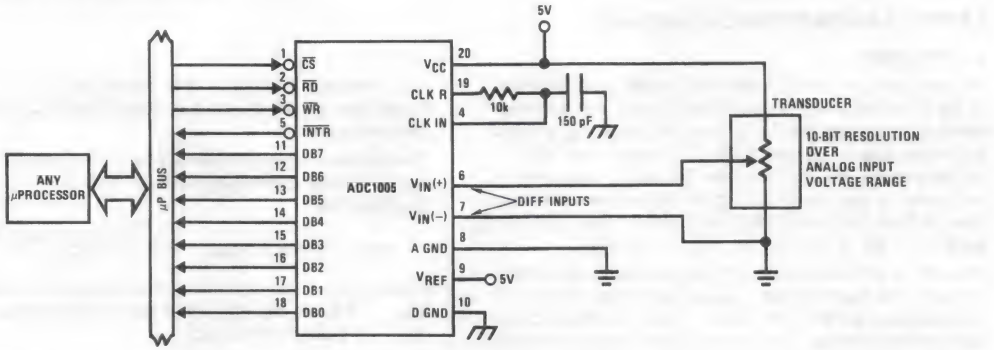
Figure 4. Zero-Shift and Span-Adjust (2V ≤ V_{IN} ≤ 5V)

Typical Applications



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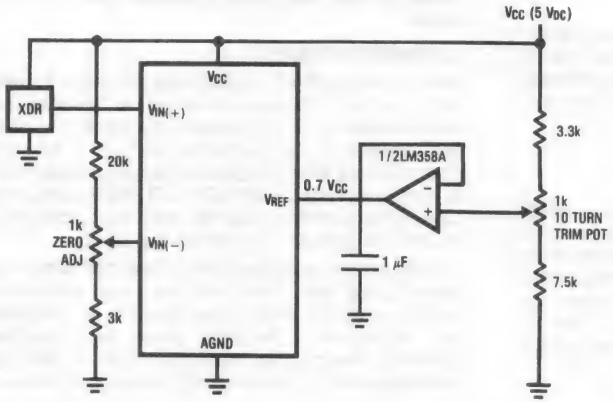
Operating with Ratiometric Transducers



$V_{IN}(-) = 0.15\ V_{CC}$

15% of $V_{CC} \leq V_{XDR} \leq 85\%$ of V_{CC}

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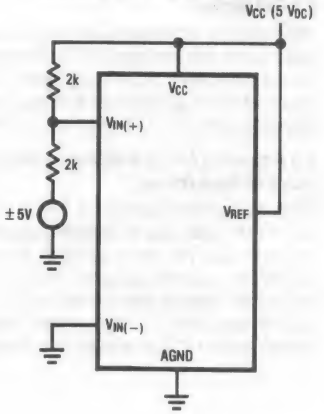
Handling ±5V Analog Inputs



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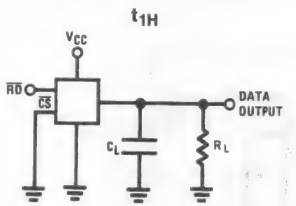
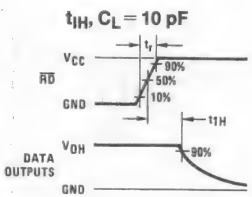
TRI-STATE Test Circuits and Waveforms

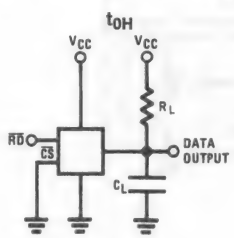


TL/H/5261–7

$t_r = 20$ ns

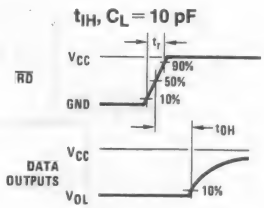


TL/H/5261–9



TL/H/5261–8

$t_r = 20$ ns



TL/H/5261–10

Ordering Information

Part Number	Package Outline	Temperature Range	Linearity Error
ADC1005BCN	N20A	0°C to +70°C	±½ LSB
ADC1025BCN	N24C		
ADC1005BCV	V20A		
ADC1025BCV	V28A		
ADC1005BCJ-1	J20A		
ADC1025BCJ-1	J24F		
ADC1005BCJ	J20A	−40°C to +85°C	
ADC1025BCJ	J24F		
ADC1005BJ	J20A	−55°C to +125°C	
ADC1025BJ	J24F		

Part Number	Package Outline	Temperature Range	Linearity Error
ADC1005CCN	N20A	0°C to +70°C	±1 LSB
ADC1025CCN	N24C		
ADC1005CCV	V20A		
ADC1025CCV	V28A		
ADC1005CCJ-1	J20A		
ADC1025CCJ-1	J24F		
ADC1005CCJ	J20A	−40°C to +85°C	
ADC1025CCJ	J24F		
ADC1005CJ	J20A	−55°C to +125°C	
ADC1025CJ	J24F		

National Semiconductor

ADC1205/ADC1225 12-Bit Plus Sign μP Compatible A/D Converters

General Description

The ADC1205 and ADC1225 are CMOS, 12-bit plus sign successive approximation A/D converters. The 24-pin ADC1205 outputs the 13-bit data result in two 8-bit bytes, formatted high-byte first with sign extended. The 28-pin ADC1225 outputs a 13-bit word in parallel for direct interface to a 16-bit data bus.

Negative numbers are represented in 2's complement data format. All digital signals are fully TTL and MOS compatible.

A unipolar input (0V to 5V) can be accommodated with a single 5V supply, while a bipolar input (−5V to +5V) requires the addition of a 5V negative supply.

The ADC1205B and ADC1225B have a maximum non-linearity over temperature of 0.012% of Full Scale, and the ADC1205C and ADC1225C have a maximum non-linearity of 0.0224% of Full Scale.

Key Specifications

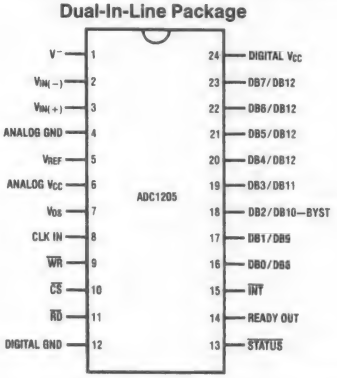
- Resolution—12 bits plus sign
- Linearity Error—±½ LSB and ±1 LSB
- Conversion Time—100 μs

Features

- Compatible with all μPs
- True differential analog voltage inputs
- 0V to 5V analog voltage range with single 5V supply
- TTL/MOS input/output compatible
- Low power—25 mW max
- Standard 24-pin or 28-pin DIP

Connection and Functional Diagrams

Dual-In-Line Package



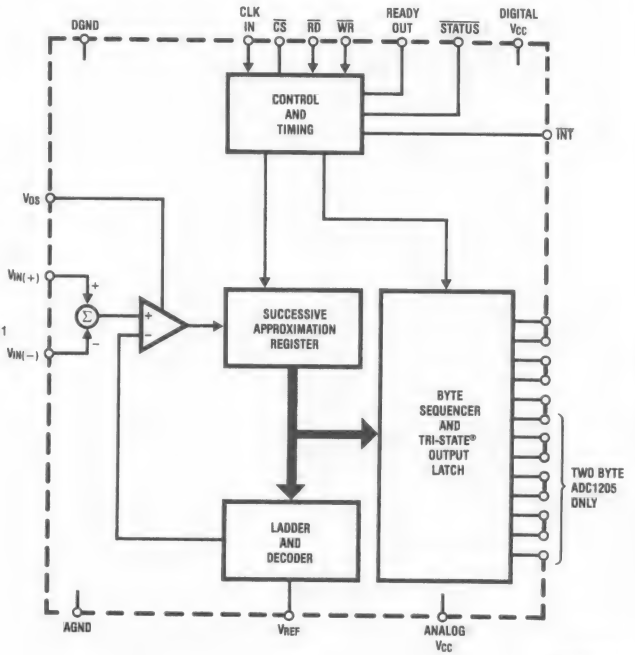
Top View

TL/H/5676-1

Dual-In-Line Package

Top View

TL/H/5676-2



TL/H/5676-3

See Ordering Information

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (DV_{CC} and AV_{CC})	6.5V
Negative Supply Voltage (V^-)	−15V to GND
Logic Control Inputs	−0.3V to +15V
Voltage at Analog Inputs $[V_{IN(+)}, V_{IN(-)}]$	$(V^-)-0.3V$ to $V_{CC}+0.3V$
Voltage at All Outputs, V_{REF}, V_{OS}	−0.3V to $(V_{CC}+0.3)V$
Input Current per Pin	±5mA
Input Current per Package	±20mA
Storage Temperature Range	−65°C to +150°C
Package Dissipation at $T_A=25°C$	875 mW
Lead Temp. (Soldering, 10 seconds)	300°C
ESD Susceptibility (Note 12)	800V

Operating Conditions (Notes 1 & 2)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
ADC1205BCJ, ADC1205CCJ	
ADC1225BCJ, ADC1225CCJ	$-40°C \le T_A \le +85°C$
ADC1205BCJ-1, ADC1205CCJ-1	
ADC1225BCJ-1, ADC1225CCJ-1	$0°C \le T_A \le 70°C$
Supply Voltage (DV_{CC} and AV_{CC})	4.5 V_{DC} to 6.0 V_{DC}
Negative Supply Voltage (V^-)	−15V to GND

Electrical Characteristics

The following specifications apply for $DV_{CC} = AV_{CC} = 5V$, $V_{REF} = 5V$, $f_{CLK} = 1.0$ MHz, $V^- = -5V$ for bipolar input range, or $V^- =$ GND for unipolar input range unless otherwise specified. Bipolar input range is defined as $-5.05V \le V_{IN(+)} \le 5.05V$; $-5.05V \le V_{IN(-)} \le 5.05V$ and $|V_{IN(+)} - V_{IN(-)}| \le 5.05V$. Unipolar input range is defined as $-0.05V \le V_{IN(+)} \le 5.05V$; $-0.05V \le V_{IN(-)} \le 5.05V$ and $|V_{IN(+)} - V_{IN(-)}| \le 5.05V$. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$ (Notes 3, 4, 5, 6, 7).

Parameter	Conditions	ADC1205BCJ, ADC1205CCJ ADC1225BCJ, ADC1225CCJ			ADC1205BCJ-1, ADC1205CCJ-1 ADC1225BCJ-1, ADC1225CCJ-1			Limit Units
		Typ (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typ (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
CONVERTER CHARACTERISTICS								
Linearity Error	Unipolar Input Range (Note 11)							
ADC1205BCJ, ADC1225BCJ			±½			±½	±½	LSB
ADC1205BCJ-1, ADC1225BCJ-1			±1					LSB
ADC1205CCJ, ADC1225CCJ						±1	±1	LSB
ADC1205CCJ-1, ADC1225CCJ-1								LSB
Unadjusted Zero Error	Unipolar Input Range		±2			±2	±2	LSB
Unadjusted Positive and Negative Full-Scale Error	Unipolar Input Range		±30			±30	±30	LSB
Negative Full-Scale Error	Unipolar Input Range, Full Scale Adj. to Zero			±½			±½	LSB
Linearity Error	Bipolar Input Range (Note 11)							
ADC1205BCJ, ADC1225BCJ			±1.5			±1.5	±1.5	LSB
ADC1205BCJ-1, ADC1225BCJ-1			±2					LSB
ADC1205CCJ, ADC1225CCJ						±2	±2	LSB
ADC1205CCJ-1, ADC1225CCJ-1								LSB
Unadjusted Zero Error	Bipolar Input Range		±2			±2	±2	LSB
Unadjusted Positive and Negative Full-Scale Error	Bipolar Input Range		±30			±30	±30	LSB
Negative Full-Scale Error	Bipolar Input Range, Full Scale Adj. to Zero		±2			±2	±2	LSB
Maximum Gain Temperature Coefficient		6		15	6		15	ppm/°C
Maximum Offset Temperature Coefficient		0.5		1.5	0.5		1.5	ppm/°C
Minimum V_{REF} Input Resistance		4.0	2		4.0	2	2	kΩ
Maximum V_{REF} Input Resistance		4.0	8		4.0	8	8	kΩ

Electrical Characteristics (Continued)

The following specifications apply for $DV_{CC} = AV_{CC} = 5V$, $V_{REF} = 5V$, $f_{CLK} = 1.0$ MHz, $V^- = -5V$ for bipolar input range, or $V^- = GND$ for unipolar input range unless otherwise specified. Bipolar input range is defined as $-5.05V \le V_{IN(+)} \le 5.05V$; $-5.05V \le V_{IN(-)} \le 5.05V$ and $|V_{IN(+)} - V_{IN(-)}| \le 5.05V$. Unipolar input range is defined as $-0.05V \le V_{IN(+)} \le 5.05V$; $-0.05V \le V_{IN(-)} \le 5.05V$ and $|V_{IN(+)} - V_{IN(-)}| \le 5.05V$. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$ (Notes 3, 4, 5, 6, 7).

Parameter	Conditions	ADC1205BCJ, ADC1205CCJ ADC1225BCJ, ADC1225CCJ			ADC1205BCJ-1, ADC1205CCJ-1 ADC1225BCJ-1, ADC1225CCJ-1			Limit Units
		Typ (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typ (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
CONVERTER CHARACTERISTICS (Continued)								
Minimum Analog Input Voltage	Unipolar Input Range		**GND-0.05**			GND-0.05	**GND-0.05**	V
	Bipolar Input Range		**$-V_{CC}-0.05$**			$-V_C-0.05$	**$-V_{CC}-0.05$**	V
Maximum Analog Input Voltage	Unipolar Input Range		**$V_{CC}+0.05$**			$V_{CC}+0.05$	**$V_{CC}+0.05$**	V
	Bipolar Input Range	**$V_{CC}+0.05$**				$V_{CC}+0.05$	**$V_{CC}+0.05$**	V
DC Common-Mode Error		±⅛	±½		±⅛	±½	±½	LSB
Power Supply Sensitivity	$AV_{CC}=DV_{CC}=5V\pm5\%$, $V^-=-5V\pm5\%$							
Zero Error			±¾			±¾	±¾	LSB
Positive and Negative Full-Scale Error			±¾			±¾	±¾	LSB
Linearity Error			±¼			±¼	±¼	LSB
DIGITAL AND DC CHARACTERISTICS								
$V_{IN(1)}$, Logical "1" Input Voltage (Min)	$V_{CC}=5.25V$, All Inputs except CLK IN		**2.0**			2.0	**2.0**	V
$V_{IN(0)}$, Logical "0" Input Voltage (Max)	$V_{CC}=4.75V$, All Inputs except CLK IN		**0.8**			0.8	**0.8**	V
$I_{IN(1)}$, Logical "1" Input Current (Max)	$V_{IN}=5V$	0.005	**1**		0.005		**1**	μA
$I_{IN(0)}$, Logical "0" Input Current (Max)	$V_{IN}=0V$	−0.005	**−1**		−0.005		**−1**	μA
V_T^+ (Min), Minimum Positive-Going Threshold Voltage	CLK IN	3.1	**2.7**		3.1	2.7	**2.7**	V
V_T^+ (Max), Maximum Positive-Going Threshold Voltage	CLK IN	3.1	**3.5**		3.1	3.5	**3.5**	V
V_T^- (Min), Minimum Negative-Going Threshold Voltage	CLK IN	1.8	**1.4**		1.8	1.4	**1.4**	V
V_T^- (Max), Maximum Negative-Going Threshold Voltage	CLK IN	1.8	**2.1**		1.8	2.1	**2.1**	V
V_H(Min), Minimum Hysteresis $[V_T^+(Min)-V_T^-(Max)]$	CLK IN	1.3	**0.6**		1.3	0.6	**0.6**	V
V_H(Max), Maximum Hysteresis $[V_T^+(Max)-V_T^-(Min)]$	CLK IN	1.3	**2.1**		1.3	2.1	**2.1**	V

Electrical Characteristics (Continued)

The following specifications apply for $DV_{CC} = AV_{CC} = 5V$, $V_{REF} = 5V$, $f_{CLK} = 1.0$ MHz, $V^- = -5V$ for bipolar input range, or $V^- = GND$ for unipolar input range unless otherwise specified. Bipolar input range is defined as $-5.05V \le V_{IN(+)} \le 5.05V$; $-5.05V \le V_{IN(-)} \le 5.05V$ and $|V_{IN(+)} - V_{IN(-)}| \le 5.05V$. Unipolar input range is defined as $-0.05V \le V_{IN(+)} \le 5.05V$; $-0.05V \le V_{IN(-)} \le 5.05V$ and $|V_{IN(+)} - V_{IN(-)}| \le 5.05V$. **Boldface limits apply from T_{MIN} to T_{MAX};** all other limits $T_A = T_J = 25°C$ (Notes 3, 4, 5, 6, 7).

Parameter	Conditions	ADC1205BCJ, ADC1205CCJ ADC1225BCJ, ADC1225CCJ			ADC1205BCJ-1, ADC1205CCJ-1 ADC1225BCJ-1, ADC1225CCJ-1			Limit Units
		Typ (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Typ (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	
DIGITAL AND DC CHARACTERISTICS (Continued)								
$V_{OUT(1)}$, Logical "1" Output Voltage (Min)	$V_{CC}=4.75V$, $I_{OUT}=-360\ \mu A$		**2.4**			2.4	**2.4**	V
	$I_{OUT}=-10\ \mu A$		**4.5**			4.5	**4.5**	V
$V_{OUT(0)}$, Logical "0" Output Voltage (Max)	$V_{CC}=4.75V$, $I_{OUT}=1.6$ mA		**0.4**			0.4	**0.4**	V
I_{OUT}, TRI-STATE Output Leakage Current (Max)	$V_{OUT}=0V$	−0.01	**−3**		−0.01	−0.3	**−3**	μA
	$V_{OUT}=5V$	0.01	**3**		0.01	0.3	**3**	μA
I_{SOURCE}, Output Source Current (Min)	$V_{OUT}=0V$	−12	**−6.0**		−12	−7.0	**−6.0**	mA
I_{SINK}, Output Sink Current (Min)	$V_{OUT}=5V$	16	**8.0**		16	9.0	**8.0**	mA
DI_{CC}, DV_{CC} Supply Current (Max)	$f_{CLK}=1$ MHz, $\overline{CS}=1$	1	**3**		1	2.5	**3**	mA
AI_{CC}, AV_{CC} Supply Current (Max)	$f_{CLK}=1$ MHz, $\overline{CS}=1$	1	**3**		1	2.5	**3**	mA
I^-, V^- Supply Current (Max)	$f_{CLK}=1$ MHz, $\overline{CS}=1$	10	**100**		10	100	**100**	μA

AC Electrical Characteristics

The following specifications apply for $DV_{CC}=AV_{CC}=5.0V$, $t_r=t_f=20$ ns and $T_A=25°C$ unless otherwise specified.

Parameter		Conditions	Typ (Note 8)	Tested Limit (Note 9)	Design Limit (Note 10)	Limit Units
f_{CLK}, Clock Frequency	MIN		1.0	0.3		MHz
	MAX		1.0	1.5		MHz
Clock Duty Cycle	MIN				40	%
	MAX				60	%
T_C, Conversion Time	MIN				108	$1/f_{CLK}$
	MAX				109	$1/f_{CLK}$
	MIN	$f_{CLK}=1.0$ MHz			108	μs
	MAX	$f_{CLK}=1.0$ MHz			109	μs
$t_{W(\overline{WR})L}$, $\overline{WR}$ Pulse Width	MAX		220		350	ns
t_{ACC}, Access Time (Delay from Falling Edge of $\overline{RD}$ to Output Data Valid) (Max)		$C_L=100$ pF	210		340	ns
t_{1H}, t_{0H}, TRI-STATE Control (Delay from Rising Edge of $\overline{RD}$ to Hi-Z State) (Max)		$R_L=2k$, $C_L=100$ pF	170		290	ns
$t_{PD(READYOUT)}$, $\overline{RD}$ or $\overline{WR}$ to READYOUT Delay (Max)			250		400	ns
$t_{PD(INT)}$, $\overline{RD}$ or $\overline{WR}$ to Reset of $\overline{INT}$ (Max)			250		400	ns

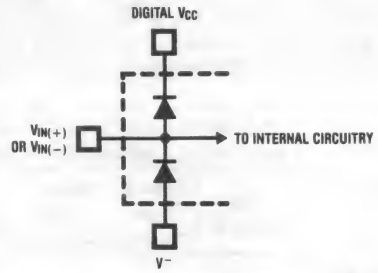
Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating ratings.

Note 2: All voltages are measured with respect to ground, unless otherwise specified.

Note 3: A parasitic zener diode exists internally from AV_{CC} and DV_{CC} to ground. This parasitic zener has a typical breakdown voltage of 7 V_{DC}.

AC Electrical Characteristics (Continued)

Note 4: Two on-chip diodes are tied to each analog input as shown below.



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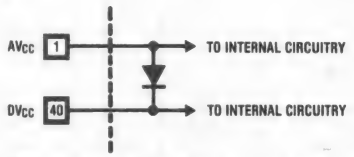
Errors in the A/D conversion can occur if these diodes are forward biased more than 50 mV. This means that if AV_{CC} and DV_{CC} are minimum (4.75 V_{DC}) and V^- is minimum ($-4.75V_{DC}$), full-scale must be $\leq 4.8V_{DC}$.

Note 5: A diode exists between analog V_{CC} and digital V_C.



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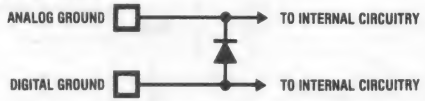
To guarantee accuracy, it is required that the AV_{CC} and DV_{CC} be connected together to a power supply with separate bypass filters at each V_{CC} pin.

Note 6: A diode exists between analog ground and digital ground.



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To guarantee accuracy, it is required that the analog ground and digital ground be connected together externally.

Note 7: Accuracy is guaranteed at $f_{CLK} = 1.0$ MHz. At higher clock frequencies accuracy may degrade.

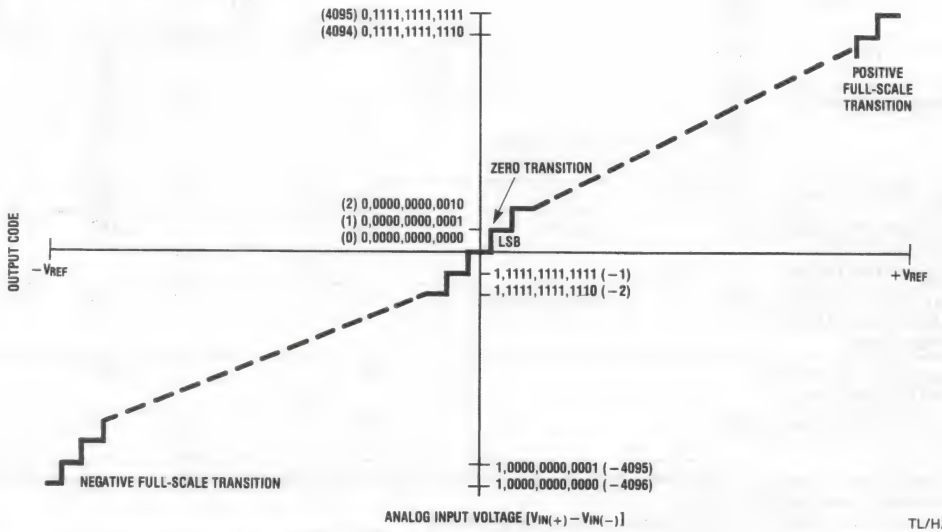
Note 8: Typicals are at 25°C and represent most likely parametric norm.

Note 9: Tested and guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 10: Guaranteed, but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Note 11: Linearity error is defined as the deviation of the analog value, expressed in LSBs, from the straight line which passes through positive full scale and zero, after adjusting zero error. (See *Figures 1b* and *1c*).

Note 12: Human body model; 100 pF discharged through a 1.5 kΩ resistor.



TL/H/5676–8

FIGURE 1a. Transfer Characteristic

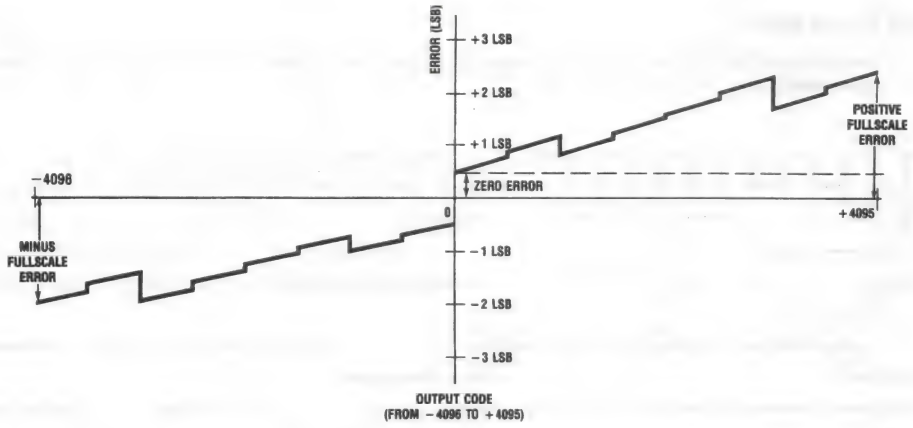


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FIGURE 1b. Simplified Error Curve vs. Output Code Without Zero and Fullscale Adjustment

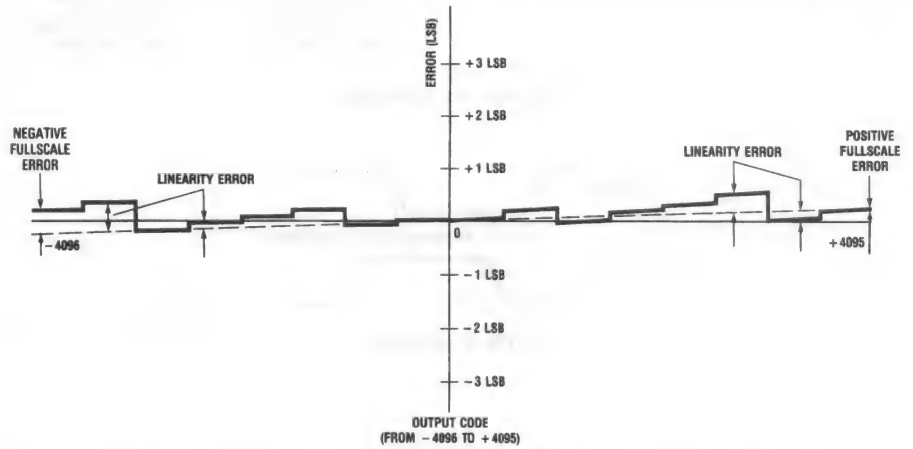

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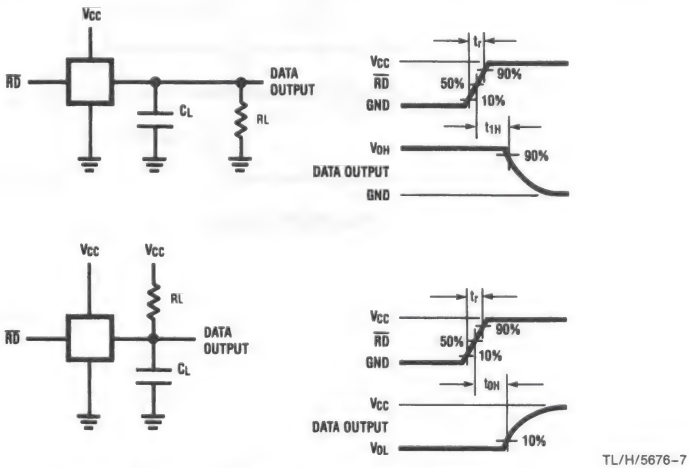
FIGURE 1c. Simplified Error Curve vs. Output Code after Zero/Fullscale Adjustment



TL/H/5676–7

FIGURE 2. TRI-STATE Test Circuits and Waveforms

Timing Diagrams

Transfer Characteristic for ADC1205 and ADC1225 Unipolar Input Range and Bipolar Input Range (digital output codes vs the difference of the analog inputs [$V_{IN(+)} - V_{IN(-)}$])

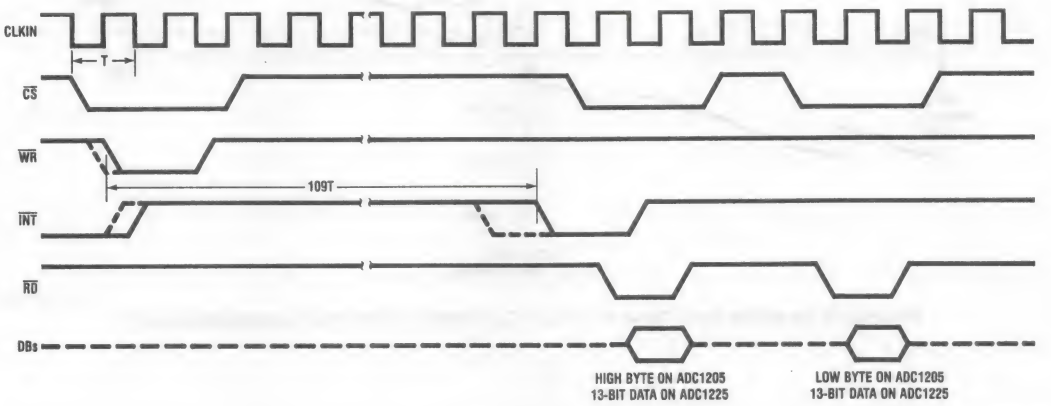


FIGURE 3. Timing Diagram

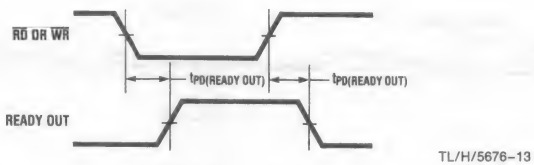


FIGURE 4. Ready Out

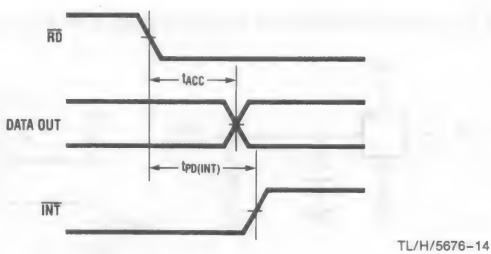


FIGURE 5. Data Out

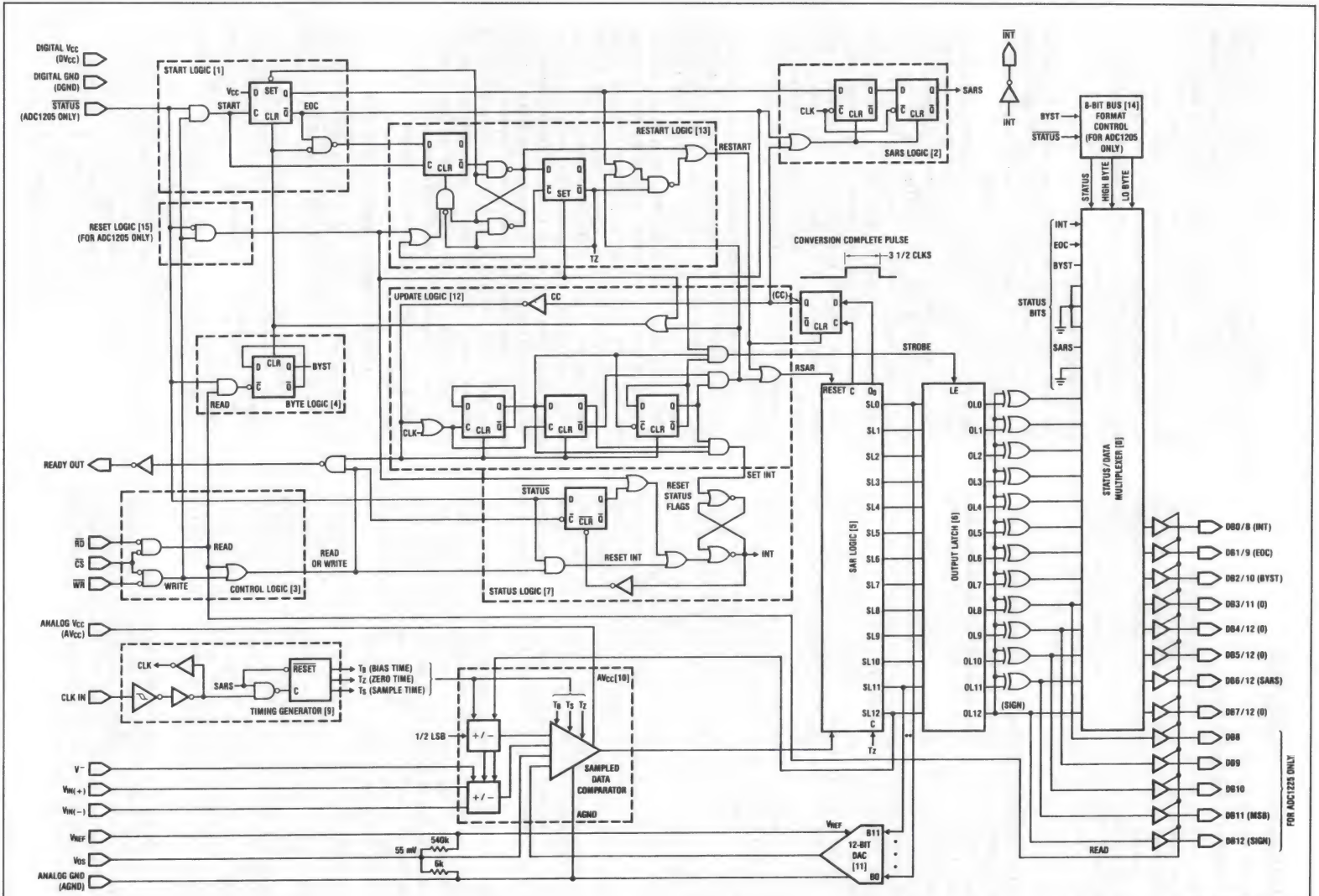


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FIGURE 6. Functional Block Diagram

Functional Description

1.0 THE A/D CONVERSION

1.1 STARTING A CONVERSION

When using the ADC1225 or ADC1205 with a microprocessor, starting an A-to-D conversion is like writing to an external memory location. The $\overline{WR}$ and $\overline{CS}$ lines are used to start the conversion. The simplified logic (*Figure 6*) shows that the falling edge of $\overline{WR}$ with $\overline{CS}$ low clocks the D-type flip-flop and initiates the conversion sequence. A new conversion can therefore be restarted before the end of the previous sequence. $\overline{INT}$ going low indicates the conversion's end.

1.2 THE CONVERSION PROCESS (Numbers designated by [ ] refer to portions of *Figure 6*.)

The SARS LOGIC [2] controls the A-to-D conversion process. When 'sars' goes high the clock (clk) is gated to the TIMING GENERATOR [9]. One of the outputs of the TIMING GENERATOR, T_Z, provides the clock for the Successive Approximation Register, SAR LOGIC [5]. The T_Z clock rate is $\frac{1}{8}$ of the CLK IN frequency.

Inputs to the 12-BIT DAC [11] and control of the SAMPLED DATA COMPARATOR [10] sign logic are provided by the SAR LOGIC. The first step in the conversion process is to set the sign to positive (logic '0') and the input of the DAC to 000 (HEX notation). If the differential input, $V_{IN(+)} - V_{IN(-)}$, is positive the sign bit will remain low. If it is negative the sign bit will be set high. Differential inputs of only a few hundred microvolts are enough to provide full logic swings at the output of the SAMPLED DATA COMPARATOR.

The sign bit indicates the polarity of the differential input. If it is set high, the negative input must have been greater than the positive input. By reversing the polarity of the differential input, $V_{IN(+)}$ and $V_{IN(-)}$ are interchanged and the DAC sees the negative input as positive. The input polarity reversal is done digitally by changing the timing on the input sampling switches of the SAMPLED DATA COMPARATOR. Thus, with almost no additional circuitry, the A/D is extended from a unipolar 12-bit to a bipolar 12-bit (12-bit plus sign) device.

After determining the input polarity, the conversion proceeds with the successive approximation process. The SAR LOGIC successively tries each bit of the 12-BIT DAC. The most significant bit (MSB), B11, has a weight of $\frac{1}{2}$ of V_{REF}. The next bit, B10, has a weight of $\frac{1}{4}$ V_{REF}. Each successive bit is reduced in weight by a factor of 2 which gives the least significant bit (LSB) a weight of 1/4096 V_{REF}.

When the MSB is tried, the comparator compares the DAC output, $V_{REF}/2$, to the analog input. If the analog input is greater than $V_{REF}/2$ the comparator tells the SAR LOGIC to set the MSB. If the analog input is less than $V_{REF}/2$ the comparator tells the SAR LOGIC to reset the MSB. On the next bit-test the DAC output will either be $\frac{3}{4}$ V_{REF} or $\frac{1}{4}$ V_{REF} depending on whether the MSB was set or not. Following this sequence through for each successive bit will approximate the analog input to within 1-bit (one part in 4096).

On completion of the LSB bit-test the conversion-complete flip-flop (CC) is set, signifying that the conversion is finished. The end-of-conversion (EOC) and interrupt ($\overline{INT}$) lines are not changed at this time. Some internal housekeeping tasks must be completed before the outside world is notified that the conversion is finished.

Setting CC enables the UPDATE LOGIC [12]. This logic controls the transfer of data from the SAR LOGIC to the OUTPUT LATCH [6] and resets the internal logic in preparation for a new conversion. This means that when EOC goes high, a new conversion can be immediately started since the internal logic has already been reset. In the same way, data is transferred to the OUTPUT LATCH prior to issuing an interrupt. This assures that data can be read immediately after $\overline{INT}$ goes low.

2.0 READING THE A/D

The ADC 1225 makes all thirteen bits of the conversion result available in parallel. Taking $\overline{CS}$ and $\overline{RD}$ low enables the TRI-STATE® output buffers. The conversion result is represented in 2's complement format.

The ADC1205 makes the conversion result available in two eight-bit bytes. The output format is 2's complement with extended sign. Data is right justified and presented high byte first. With $\overline{CS}$ low and $\overline{STATUS}$ high, the high byte (DB12–DB8) will be enabled on the output buffers the first time $\overline{RD}$ goes low. When $\overline{RD}$ goes low a second time, the low byte (DB7–DB0) will be enabled. On each read operation, the 'byst' flip-flop is toggled so that on successive reads alternate bytes will be available on the outputs. The 'byst' flip-flop is always reset to the high byte at the end of a conversion. Table 1 below shows the data bit locations on the ADC1205.

The ADC1205's $\overline{STATUS}$ pin makes it possible to read the conversion status and the state of the 'byst' flip-flop. With $\overline{RD}$, $\overline{STATUS}$ and $\overline{CS}$ low, this information appears on the data bus. The 'byst' status appears on pin 18 (DB2/DB10). A low output on pin 18 indicates that the next data read will be the high byte. A high output indicates that the next data read will be the low byte. A high status bit on pin 22 (DB6/DB12) indicates that the conversion is in progress. A high output appears on pin 17 (DB1/DB9) when the conversion is completed and the data has been transferred to the output latch. A high output on pin 16 (DB0/DB8) indicates that the conversion has been completed and the data is ready to read. This status bit is reset when a new conversion is initiated, data is read, or status is read. When reading status or a conversion result, $\overline{STATUS}$ should always change states at least 600 ns before $\overline{RD}$ goes low. If the conversion status information is not needed, the $\overline{STATUS}$ pin should be hardwired to V^+. Table 2 summarizes the meanings of the four status bits.

TABLE I. Data Bit Locations, ADC1205

HIGH BYTE	DB12	DB12	DB12	DB12	DB11	DB10	DB9	DB8
LOW BYTE	DB7	DB6	DB5	DB4	DB3	DB2	DB1	DB0

TABLE II. Status Bit Locations and Meanings

Status Bit Location	Status Bit	Meaning	Condition to Clear Status Bit
DB6	SARS	"High" indicates that the conversion is in progress	
DB2	BYST	"Low" indicates that the next data read is the high byte. "High" indicates that the next data read is the low byte	Status write or toggle it with data read

Functional Description (Continued)

TABLE II. Status Bit Locations and Meanings (Continued)

Status Bit Location	Status Bit	Meaning	Condition to Clear Status Bit
DB1	EOC	"High" indicates that the conversion is completed and data is transferred to the output latch.	
DB0	INT	"High" indicates that it is the end of the conversion and the data is ready to read	Data read or status read or status write

3.0 INTERFACE

3.1 RESET OF INTERRUPT

$\overline{INT}$ goes low at the end of the conversion and indicates that data is transferred to the output latch. By reading data, $\overline{INT}$ will be reset to high on the leading edge of the first read ($\overline{RD}$ going low). $\overline{INT}$ is also reset on the leading (falling) edge of $\overline{WR}$ when starting a conversion.

3.2 READY OUT

To simplify the hardware connection to high speed microprocessors, a READY OUT line is provided. This allows the A-to-D to insert a wait state in the μP's read cycle. The equivalent circuit and the timing diagram for READY OUT is shown in *Figures 7* and *8*.

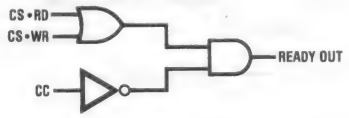

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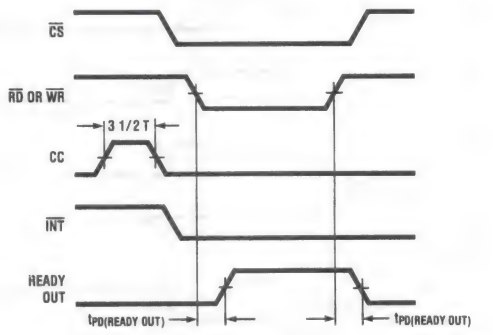
FIGURE 7. READY OUT Equivalent Circuit



TL/H/5676–10

FIGURE 8. READY OUT Timing Diagram

3.3 RESETTING THE A/D

All the internal logic can be reset, which will abort any conversion in process and reset the status bits. The reset function is achieved by performing a status write ($\overline{CS}$, $\overline{WR}$ and $\overline{STATUS}$ are low).

3.4 ADDITIONAL TIMING AND INTERFACE OPTIONS

ADC1225

1. $\overline{WR}$ and $\overline{RD}$ can be tied together with $\overline{CS}$ low continuously or strobed. The previous conversion's data will be available when the $\overline{WR}$ and $\overline{RD}$ are low as shown below.

One drawback is that, since the conversion is started on the falling edge and the data read on the rising edge of $\overline{WR}/\overline{RD}$, the first data access will have erroneous information depending on the power-up state of the internal output latches.

If the $\overline{WR}/\overline{RD}$ strobe is longer than the conversion time, $\overline{INTR}$ will never go low to signal the end of a conversion. The conversion will be completed and the output latches will be updated. In this case the READY OUT signal can be used to sense the end of the conversion since it will go low when the output latches are being updated.

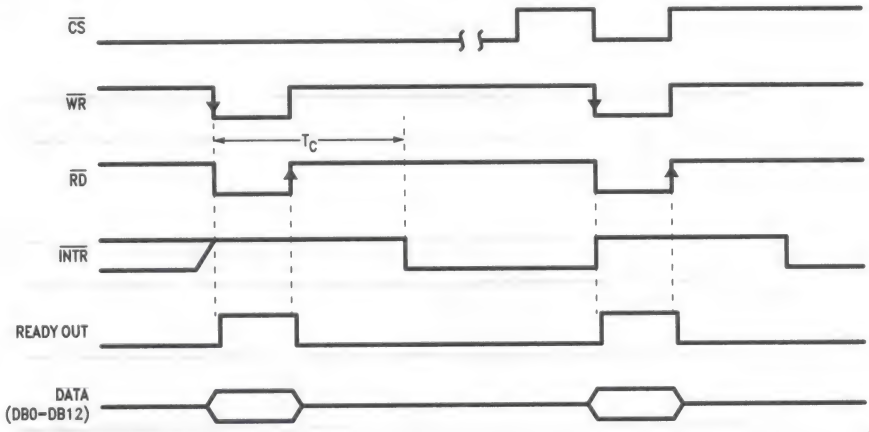


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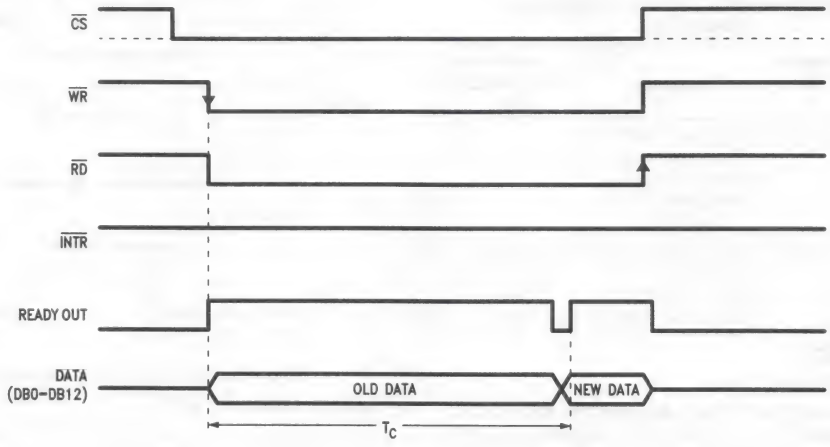
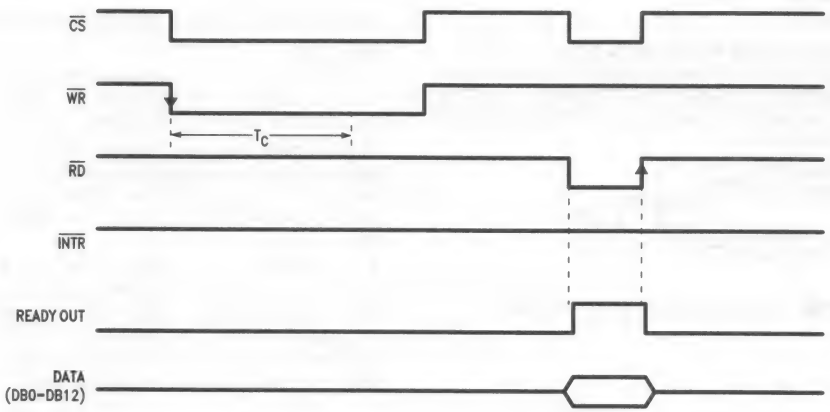
FIGURE 9

Functional Description (Continued)



TL/H/5676–25

FIGURE 10



TL/H/5676–26

FIGURE 11

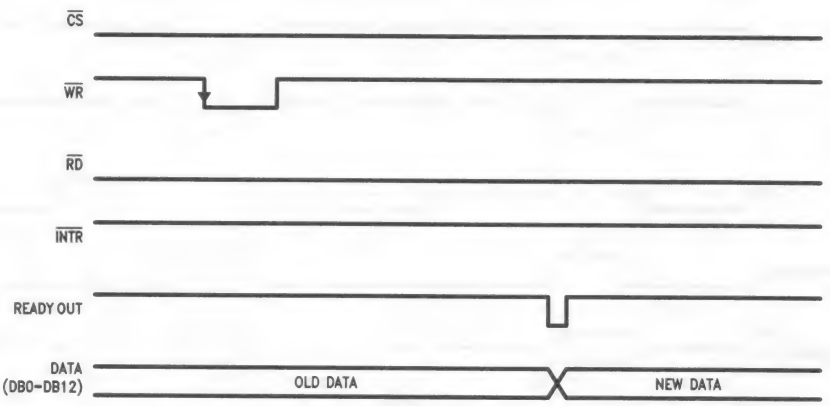


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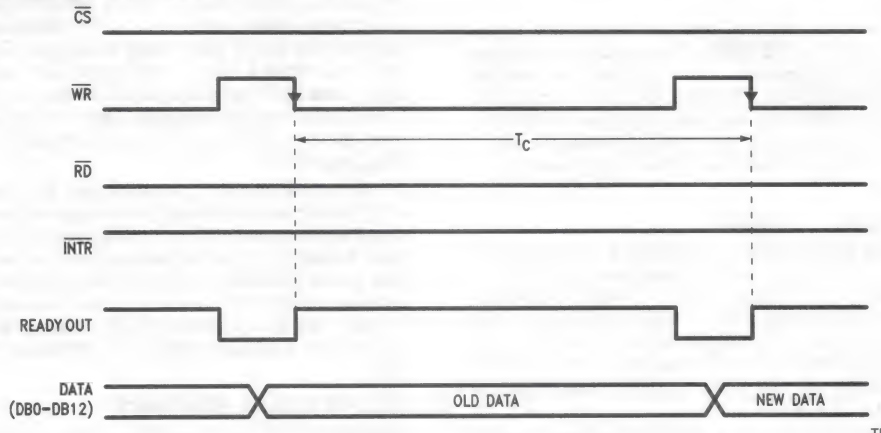
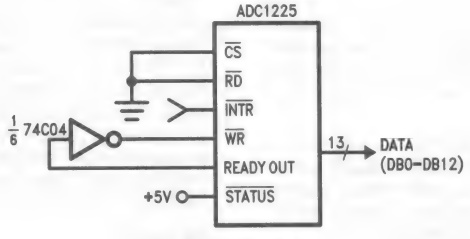
FIGURE 12

Functional Description (Continued)



TL/H/5676–28



TL/H/5676–29

FIGURE 13

When using this method of conversion only one strobe is necessary and the rising edge of $\overline{WR}$/$\overline{RD}$ can be used to read the current conversion results. These methods reduce the throughput time of the conversion since the $\overline{RD}$ and $\overline{WR}$ cycles are combined.

2. With the standard timing $\overline{WR}$ pulse width longer than the conversion time a conversion is completed but the $\overline{INTR}$ will never go low to signal the end of a conversion. The output latches will be updated and valid information will be available when the $\overline{RD}$ cycle is accomplished.

3. Tying $\overline{CS}$ and $\overline{RD}$ low continuously and strobing $\overline{WR}$ to initiate a conversion will also yield valid data. The $\overline{INTR}$ will never go low to signal the end of a conversion and the digital outputs will always be enabled, so using $\overline{INTR}$ to strobe the $\overline{WR}$ line for a continuous conversion cannot be done with this part.

A simple stand-alone circuit can be accomplished by driving $\overline{WR}$ with the inverse of the READY OUT signal using a simple inverter as shown below.

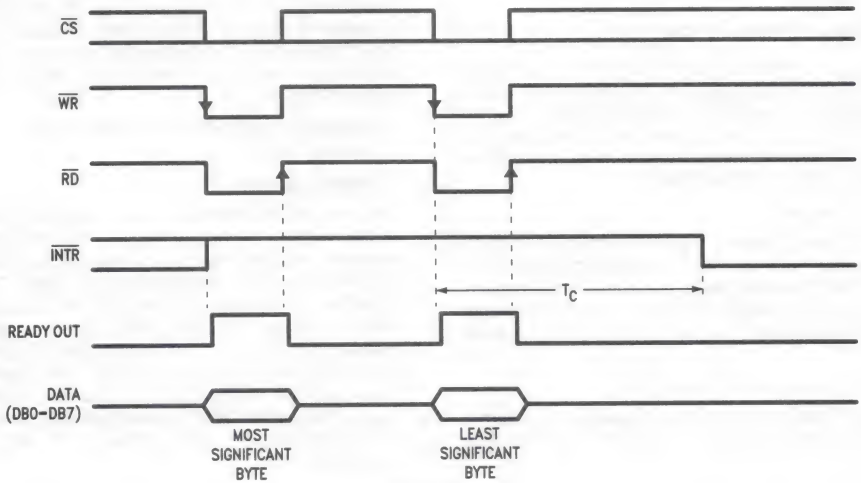


FIGURE 14

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Functional Description (Continued)

ADC1205

Case 1 would be the only one that would appy to the ADC1205 since two $\overline{RD}$ strobes are necessary to retrieve the 13 bits of information on the 8 bit data bus. Simultaneously strobing $\overline{WR}$ and $\overline{RD}$ low will enable the most significant byte on DB0–DB7 and start a conversion. Pulsing $\overline{WR}/\overline{RD}$ low before the end of this conversion will enable the least significant byte of data on the outputs and restart a conversion.

4.0 REFERENCE VOLTAGE

The voltage applied to the reference input of the converter defines the voltage span of the analog inputs (the difference between $V_{IN(+)}$ and $V_{IN(-)}$, over which 4096 positive output codes and 4096 negative output codes exist. The A-to-D can be used in either ratiometric or absolute reference applications. V_{REF} must be connected to a voltage source capable of driving the reference input resistance (typically 4 kΩ).

In a ratiometric system, the analog input voltage is proportional to the voltage used for the A/D reference. When this voltage is the system power supply, the V_{REF} pin can be tied to V_{CC}. This technique relaxes the stability requirement of the system reference as the analog input and A/D reference move together maintaining the same output code for a given input condition.

For absolute accuracy, where the analog input varies between very specific voltage limits, the reference pin can be biased with a time and temperature stable voltage source. In general, the magnitude of the reference voltage will require an initial adjustment to null out full-scale errors.

5.0 THE ANALOG INPUTS

5.1 DIFFERENTIAL VOLTAGE INPUTS AND COMMON MODE REJECTION

The differential inputs of the ADC1225 and ADC1205 actually reduce the effects of common-mode input noise, i.e., signals common to both $V_{IN(+)}$ and $V_{IN(-)}$ inputs (60 Hz is most typical). The time interval between sampling the "+" and "−" input is 4 clock periods. Therefore, a change in the common-mode voltage during this short time interval may cause conversion errors. For a sinusoidal common-mode signal the error would be:

$$V_{ERROR(MAX)} = V_{PEAK}\,(2\pi\, f_{CM})\,\frac{4}{f_{CLK}}$$

where f_{CM} is the frequency of the common-mode signal, V_{PEAK} is its peak voltage value and f_{CLK} is the converter's clock frequency. In most cases V_{ERROR} will not be significant. For a 60 Hz common-mode signal to generate a ¼ LSB error (300 μV) with the converter running at 1 MHz its peak value would have to be 200mV.

5.2 INPUT CURRENT

Due to the sampling nature of the analog inputs, short duration spikes of current enter the "+" input and exit the "−" input at the leading clock edges during the actual conversion. These currents decay rapidly and do not cause errors as the internal comparator is strobed at the end of a clock period.

5.3 INPUT BYPASS CAPACITORS

Bypass capacitors at the inputs will average the current spikes mentioned in 5.2 and cause a DC current to flow through the output resistance of the analog signal source. This charge pumping action is worse for continuous conversions with the $V_{IN(+)}$ input voltage at full-scale. For continuous conversions with a 1 MHz clock frequency and the $V_{IN(+)}$ input at 5V, the average input current is approximately 5 μA. For this reason bypass capacitors should not be used at the analog inputs for high resistance sources (R_{SOURCE} 100 Ω).

If input bypass capacitors are necessary for noise filtering and high source resistance is desirable to minimize capacitor size, the detrimental effects of the voltage drop across this input resistance, due to the average value of the input current, can be minimized with a full-scale adjustment while the given source resistance and input bypass capacitor are both in place. This is effective because the average value of the input current is a linear function of the differential input voltage.

5.4 INPUT SOURCE RESISTANCE

Large values of source resistance where an input bypass capacitor is not used, will not cause errors as the input currents settle out prior to the comparison time. If a low pass filter is required in the system, use a low valued series resistor ($R \leq 100\ \Omega$) for a passive RC section or add an op amp RC active low pass filter. For low source resistance applications, ($R_{SOURCE} \leq 100\ \Omega$) a 0.001 μF bypass capacitor at the inputs will prevent pickup due to series lead inductance of a long wire. A 100 Ω series resistor can be used to isolate this capacitor – both the R and C are placed outside the feedback loop – from the output of an op amp, if used.

5.5 NOISE

The leads to the analog inputs should be kept as short as possible to minimize input noise coupling. Both noise and undesired digital clock coupling to these inputs can cause errors. Input filtering can be used to reduce the effects of these sources, but careful note should be taken of sections 5.3 and 5.4 if this route is taken.

6.0 POWER SUPPLIES

Noise spikes on the V_{CC} supply line can cause conversion errors as the comparator will respond to this noise. Low inductance tantalum capacitors of 1 μF or greater are recommended for supply bypassing. Separate bypass caps should be placed close to the DV_{CC} and AV_{CC} pins. If an unregulated voltage source is available in the system, a separate LM340LAZ-5.0 voltage regulator for the A-to-D's V_{CC} (and other analog circuitry) will greatly reduce digital noise on the supply line.

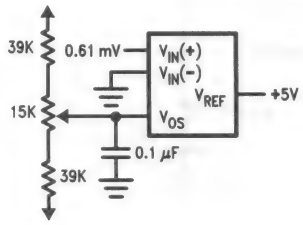
7.0 ERRORS AND REFERENCE VOLTAGE ADJUSTMENTS

7.1 ZERO ADJUST

The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the $V_{IN(-)}$ input and applying a small magnitude positive voltage to the $V_{IN(+)}$ input. Zero error is the difference between the actual DC input voltage necessary to just cause an output digital code transition from all zeroes to 0,0000,0000,0001 and the ideal ½ LSB value (½ LSB = 0.61 mV for $V_{REF} = 5\ V_{DC}$). Zero error can be adjusted as shown in *Figure 15.* $V_{IN(+)}$ is forced to 0.61 mV, and $V_{IN(-)}$ is forced to 0V. The potentiometer is adjusted until the digital output code changes from all zeroes to 0,000,0000,0001.

Functional Description (Continued)

A simpler, although slightly less accurate, approach is to ground $V_{IN(+)}$ and $V_{IN(-)}$, and adjust for all zeros at the output. Error will be well under ½ LSB if the adjustment is done so that the potentiometer is "centered" within the 0,000,000 range. A positive voltage at the V_{OS} input will reduce the output code. The adjustment range is +4 to −30 LSB.



TL/H/5676–11

FIGURE 15. Zero Adjust Circuit

7.2 POSITIVE AND NEGATIVE FULL-SCALE ADJUSTMENT

Unipolar Inputs

Apply a differential input voltage which is 1.5 LSB below the desired analog full-scale voltage (V_F) and adjust the magnitude of the V_{REF} input so that the output code is just changing from 0,1111,1111,1110 to 0,1111,1111,1111.

Bipolar Inputs

Do the same procedure outlined above for the unipolar case and then change the differential input voltage so that the digital output code is just changing from 1,0000,0000,0001 to 1,0000,0000,0000. Record the differential input voltage, V_X. the ideal differential input voltage for that transition should be;

$$\left(-V_F + \frac{V_F}{8192}\right)$$

Calculate the difference between Vx and the ideal voltage;

$$\Delta = V_X - \left(-V_F + \frac{V_F}{8192}\right)$$

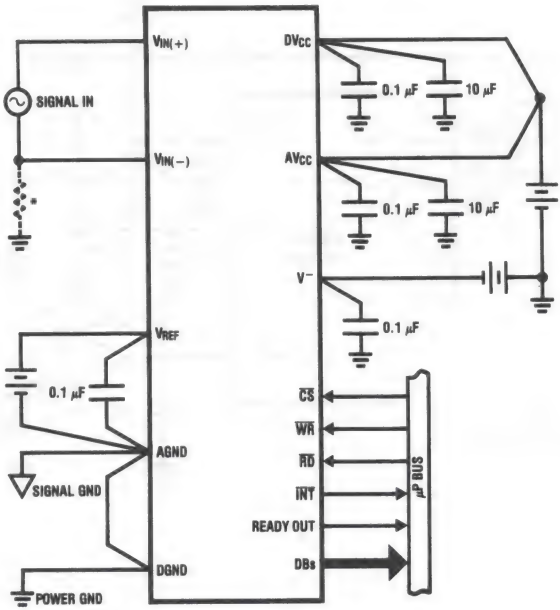
Then apply a differential input voltage of;

$$\left(V_X - \frac{\Delta}{2}\right)$$

and adjust the magnitude of V_{REF} so the digital output code is just changing from 1,0000,0000,0001 to 1,0000,0000,0000. That will obtain the positive and negative full-scale transition with symmetrical minimum error.

Typical Applications

*Input must have some current return path to signal ground



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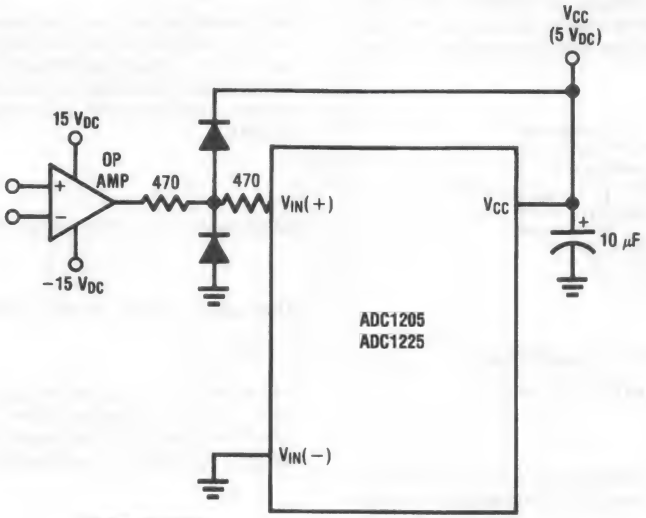
Typical Applications (Continued)

Protecting the Input



Diodes are 1N914

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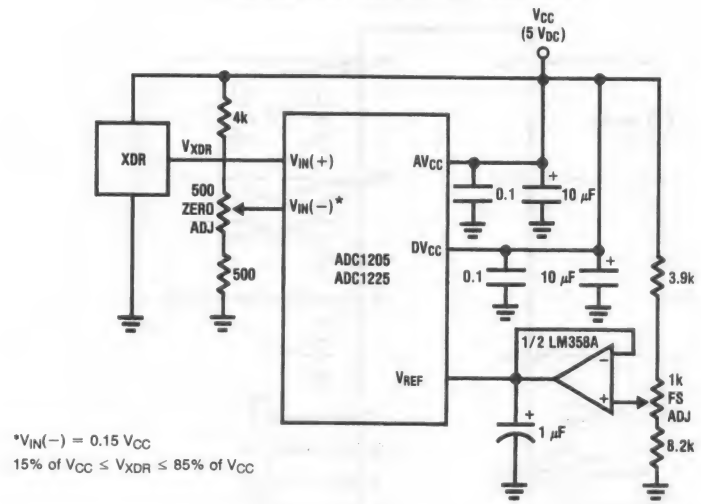
Operating with Ratiometric Transducers



*VIN(−) = 0.15 VCC

15% of VCC ≤ VXDR ≤ 85% of VCC

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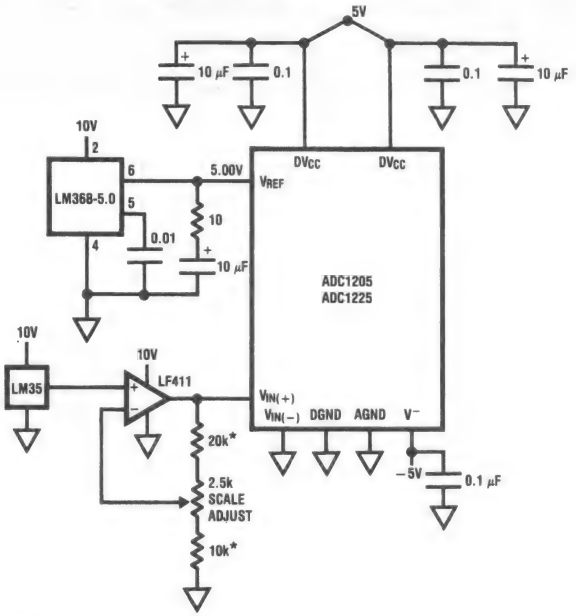
Typical Applications (Continued)

Bipolar Input Temperature Converter



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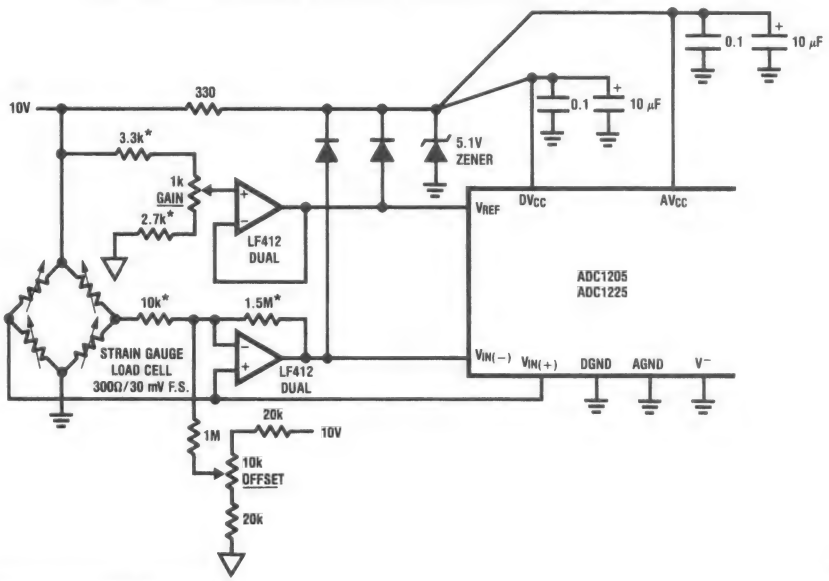
+150 to −55°C with 0.04°C resolution

Note: * resistors are 1% metal film types

Strain Gauge Converter with .025% Resolution and Single Power Supply



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Note: 1)* resistors are 1% metal film types

2) LF412 power +10V and ground

Ordering Information

Temperature Range		0°C to 70°C		−40°C to +85°C	
Non-Linearity	0.012%	ADC1205BCJ-1	ADC1225BCJ-1	ADC1205BCJ	ADC1225BCJ
	0.024%	ADC1205CCJ-1	ADC1225CCJ-1	ADC1205CCJ	ADC1225CCJ
Package Outline		J24A	J28A	J24A	J28A

National Semiconductor

ADC1210/ADC1211 12-Bit CMOS A/D Converters

General Description

The ADC1210, ADC1211 are low power, medium speed, 12-bit successive approximation, analog-to-digital converters. The devices are complete converters requiring only the application of a reference voltage and a clock for operation. Included within the device are the successive approximation logic, CMOS analog switches, precision laser trimmed thin film R-2R ladder network and FET input comparator.

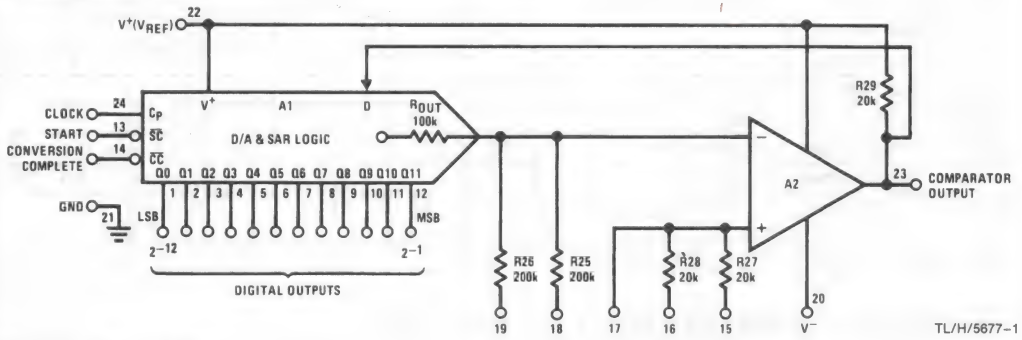
The ADC1210 offers 12-bit resolution and 12-bit accuracy, and the ADC1211 offers 12-bit resolution with 10-bit accuracy. The inverted binary outputs are directly compatible with CMOS logic. The ADC1210, ADC1211 will operate over a wide supply range, convert both bipolar and unipolar analog inputs, and operate in either a continuous conversion mode or logic-controlled START-STOP conversion mode. The devices are capable of making a 12-bit conversion in 100 μs typ, and can be connected to convert 10 bits in 30 μs.

Both devices are available in military and industrial temperature ranges.

Features

- 12-bit resolution
- $\pm\frac{3}{4}$ LSB or ± 2 LSB nonlinearity
- Single $+5$V to ± 15V supply range
- 100 μs 12-bit, 30 μs 10-bit conversion rate
- CMOS compatible outputs
- Bipolar or unipolar analog inputs
- 200 kΩ analog input impedance

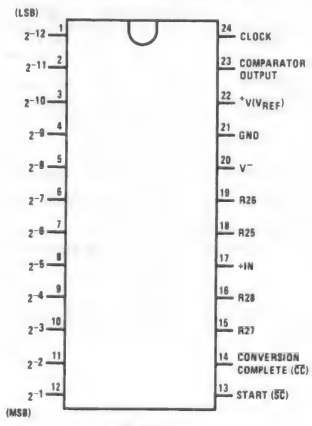
Block Diagram



TL/H/5677–1

Connection Diagram

Dual-In-Line Package



TL/H/5677–2

Order Number ADC1210HD, ADC1210HCD, ADC1211HD, ADC1211HCD
See NS Package D24D

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Maximum Reference Supply Voltage (V^+)	16V
Maximum Negative Supply Voltage (V^-)	−20V
Voltage At Any Logic Pin	$V^+ + 0.3V$
Analog Input Voltage	±15V
Maximum Digital Output Current	±10 mA
Maximum Comparator Output Current	50 mA
Comparator Output Short-Circuit Duration	5 Seconds
Power Dissipation	See Curves
Operating Temperature Range	
ADC1210HD, ADC1211HD	−55°C to +125°C
ADC1210HCD, ADC1211HCD	−25°C to +85°C
Storage Temperature Range	−65°C to +150°C
Lead Temperature (Soldering, 10 seconds)	300°C
ESD Susceptibility (Note 4)	TBD V

DC Electrical Characteristics (Notes 1 and 2)

Parameter	Conditions	ADC1210			ADC1211			Units
		Min	Typ	Max	Min	Typ	Max	
Resolution		12			12			Bits
Linearity Error	(Note 3) $f_{CLK}=65$ kHz, $T_A=25°C$ $f_{CLK}=65$ kHz			±0.0183 ±0.0366			±0.0488	% FS % FS
Full Scale Error	$T_A=25°C$, Unadjusted			0.20			0.50	% FS
Zero Scale Error	$T_A=25°C$, Unadjusted			0.20			0.50	% FS
Quantization Error				±1/2			±1/2	LSB
Input Resistor Values	R27, R28		20			20		kΩ
Input Resistor Values	R25, R26		200			200		kΩ
Input Resistor Ratios	R25/R26, R27/R28			0.8			0.8	%
Logic "1" Input Voltage		8			8			V
Logic "0" Input Voltage				2			2	V
Logic "1" Input Current	$V_{IN}=10.24V$			1			1	μA
Logic "0" Input Current	$V_{IN}=0V$			−1			−1	μA
Logic "1" Output Voltage	$I_{OUT}\leq -1$ μA	9.2			9.2			V
Logic "0" Output Voltage	$I_{OUT}\leq 1$ μA			0.5			0.5	V
Positive Supply Current	$V^+=15V$, $f_{CLK}=65$ kHz, $T_A=25°C$		5	8		5	8	mA
Negative Supply Current	$V^-=-15V$, $T_A=25°C$		4	6		4	6	mA

AC Electrical Characteristics $T_A=25°C$, (Notes 1 and 2)

Parameter	Conditions	Min	Typ	Max	Units
Conversion Time			100	200	μs
Maximum Clock Frequency			130	65	kHz
Clock Pulse Width		100	50		ns
Propagation Delay From Clock to Data Output	$t_r\leq t_f\leq 10$ ns		60	150	ns
(Q0 to Q11)					
Propagation Delay from Clock to Conversion	$t_r\leq t_f\leq 10$ ns		60	150	ns
Complete					
Clock Rise and Fall Time				5	μs
Input Capacitance			10		pF
Start Conversion Set-Up Time		30			ns

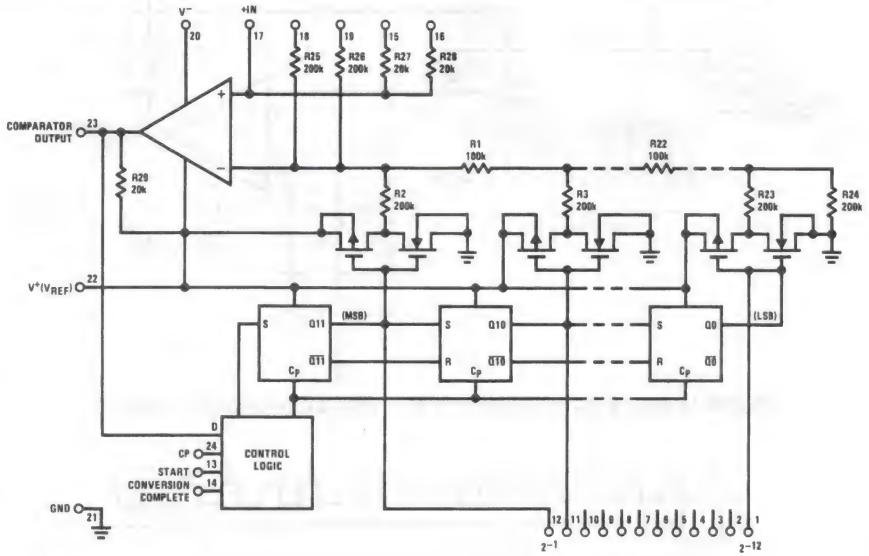
Note 1: Unless otherwise noted, these specifications apply for $V^+=10.240V$, $V^-=-15V$, over the temperature range −55°C to +125°C for the ADC1210HD, ADC1211HD, and −25°C to +85°C for the ADC1210HCD, ADC1211HCD.

Note 2: All typical values are for $T_A=25°C$.

Note 3: Unless otherwise noted, this specification applies over the temperature range −25°C to +85°C. Provision is made to adjust zero scale error to 0V and full-scale to 10.2375V during testing. Standard linearity test circuit is shown in *Figure 5a*.

Note 4: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

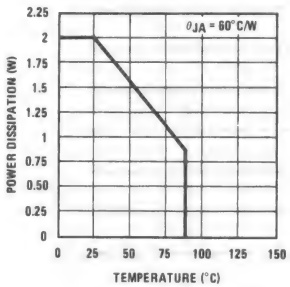
Schematic Diagram



TL/H/5677–3

Note: 3 bits shown for clarity

Power Dissipation vs Temperature



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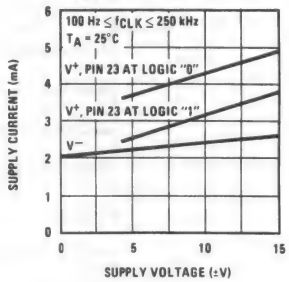
Supply Current vs Supply Voltage



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1.0 THEORY OF OPERATION

The ADC1210, ADC1211 are successive approximation analog-to-digital converters, i.e., the conversion takes place 1 bit at a time by comparing the output of the internal D/A to the (unknown) input voltage. The START input (pin 13), when taken low, causes the register to reset synchronously on the next CLOCK low-to-high transition. The MSB, Q11 is set to the low state, and the remaining bits, Q0 through Q10, will be set to the high state. The register will remain in this state until the $\overline{SC}$ input is taken high. When START goes high, the conversion will begin on the low-to-high transition of the CLOCK pulse. Q11 will then assume the state of pin 23. If pin 23 is high, Q11 will be high; if pin 23 is low, Q11 will remain low. At the same time, the next bit Q10 is set low. All remaining bits, Q0–Q9 will remain unchanged (high). This process will continue until the LSB (Q0) is found. When the conversion process is completed, it is indicated by CONVERSION COMPLETE ($\overline{CC}$) (pin 14) going low. The logic levels at the data output pins (pins 1–12) are the complemented-binary representation of the converted analog signal with Q11 being the MSB and Q0 being the LSB. The register will remain in the above state until the $\overline{SC}$ is again taken low.

An application example is shown in *Figure 1*. In this case, a 0 to −10.2375V input is being converted using the ADC1210 with $V^+ = 10.240V$, $V^- = -15V$. *Figure 1b* is the timing diagram for full scale input. *Figure 1c* is the timing diagram for zero scale input, *Figure 1d* is the timing diagram for −3.4125V input (010101010101 = output).

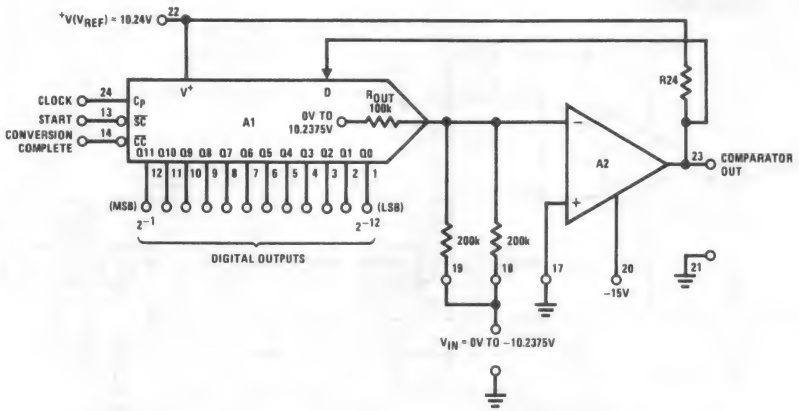

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FIGURE 1a. ADC1210 Connected for 0V to −10.2375V (Natural Binary Output)

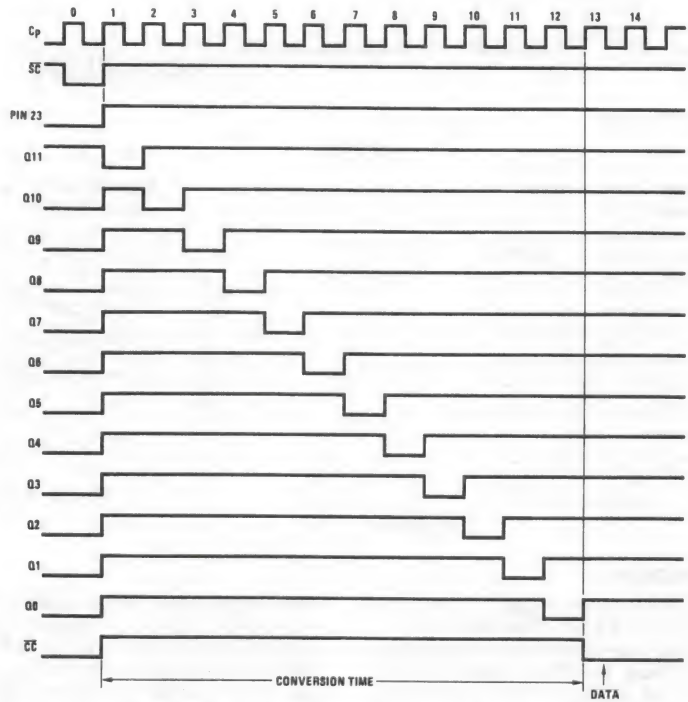

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FIGURE 1b. Timing Diagram for V_{IN} = Full Scale Input

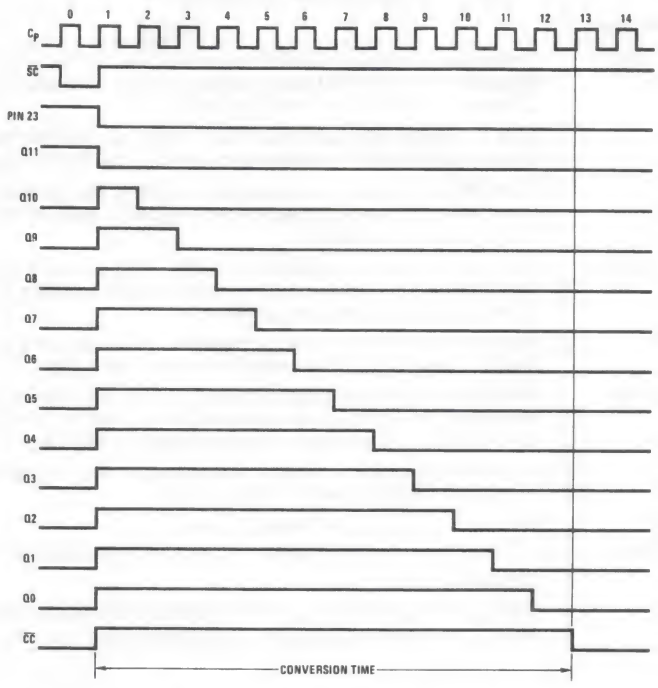

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FIGURE 1c. Timing Diagram for V_{IN} = Zero Scale

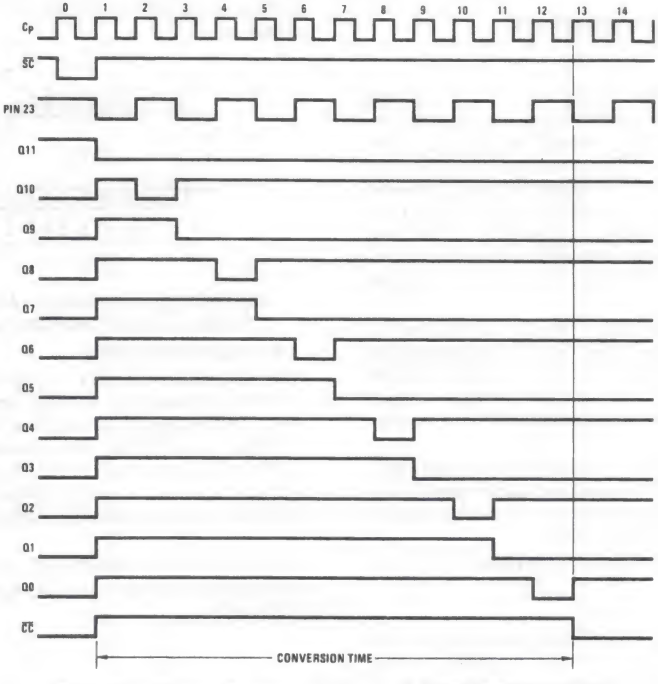

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FIGURE 1d. Timing Diagram for V_{IN} = −3.4125V (010101010101)

TABLE 1. Pin Assignments and Explanations

Pin Number	Mnemonic	Function
1–12	Q11–Q0	Digital (data) output pins. This information is a parallel 12-bit complemented binary representation of the converted analog signal. All data is valid when "Conversion Complete" goes low. Logic levels are ground and V^+.
13	$\overline{SC}$	Start Conversion is a logic input which causes synchronous reset of the successive approximation register and initiates conversion. Logic levels are ground and V^+.
14	$\overline{CC}$	"Conversion Complete" is a digital output signal which indicates the status of the converter. When $\overline{CC}$ is high, conversion is taking place, when low conversion is completed. Logic levels are ground and V^+.
15, 16	R27, R28	R27 and R28 are two application resistors connected to the comparator non-inverting input. The resistors may be used in various modes of operation. Their nominal values are 20 kΩ each. See Applications section.
17	+IN	Non-inverting input of the analog comparator. This node is used in various configurations and for compensation of the loop. See Applications section.
18, 19	R25, R26	R25 and R26 are two application resistors that are tied internally to the inverting input of the comparator. Their nominal values are 200 kΩ each. See Applications section. The R-2R ladder network will have the same temperature coefficient as these resistors.
20	V^-	Negative supply voltage for bias of the analog comparator. Optionally may be grounded or operated with voltages to −20V.
21	GND	Ground for both digital and analog signals.
22	V^+ (V_{REF})	V^+ sets both maximum full scale and input and output logic levels.
23	CO	Comparator output.
24	C_P	Clock is an input which causes the successive approximation (shift) register to advance through the conversion sequence. Logic levels are ground and V^+.

2.0 APPLICATIONS

2.1 Power Supply Considerations and Decoupling

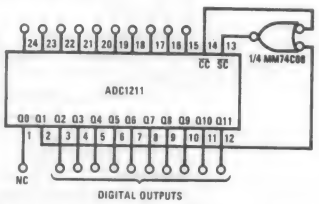
Pin 22 is both the positive supply and voltage reference input to the ADC1210, ADC1211. The magnitude of V^+ determines the input logic "1" threshold and the output voltage from the CMOS SAR. The device will operate over a range of V^+ from 5V to 15V. However, in order to preserve 12-bit accuracy, V^+ should be well regulated (0.01%) and isolated from external switching transients. It is therefore recommended that pin 22 be decoupled with a 4.7 μF tantalum capacitor in parallel with a 0.1 μF ceramic disc capacitor.

The V^- supply (pin 20) provides negative bias for the FET comparator. Although pin 20 may be grounded in some applications, it must be at least 2V more negative than the most negative analog input signal. When a negative supply is used, pin 20 should also be bypassed with 4.7 μF in parallel with 0.1 μF.

Grounding and circuit layout are extremely important in preserving 12-bit accuracy. The user is advised to employ separate digital and analog returns, and to make these PC board traces as "heavy" as practical.

2.2 Short Cycle for Improved Conversion Time (Figure 2)

The ADC1210, ADC1211 counting sequence may be truncated to decrease conversion time. For example, when using the ADC1211, 2 clock intervals may be "saved" if 10-bit conversion accuracy is taking place. The Q2 output should be "OR'd" with CONVERSION COMPLETE ($\overline{CC}$) in order to ensure that the register does not lock-up upon power turn-on.



TL/H/5677–10

FIGURE 2. Short Cycling the ADC1211 to Improve 10-Bit Conversion Time (Continuous Conversion)

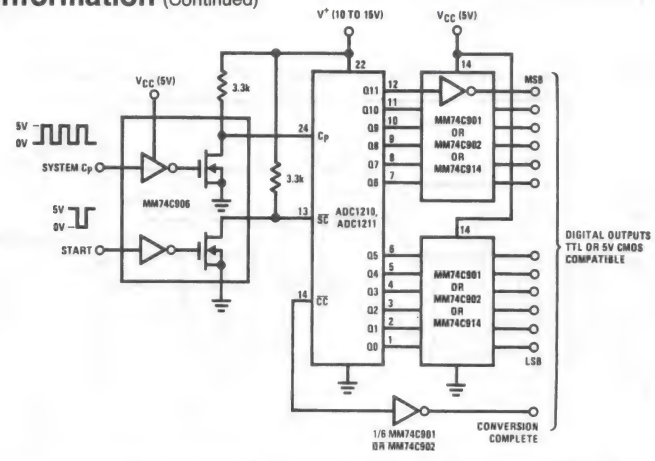
2.3 Logic Compatibility

The ADC1210, ADC1211 is intended to interface with CMOS logic levels: i.e., the logic inputs and outputs are directly compatible with series 54C/74C and CD4000 family of logic components. The outputs of the ADC1210, ADC1211 will not drive LPTTL, TTL or PMOS logic directly without degrading accuracy. Various recommended interface techniques are shown in *Figures 3 and 4*.

2.4 Operating Configurations

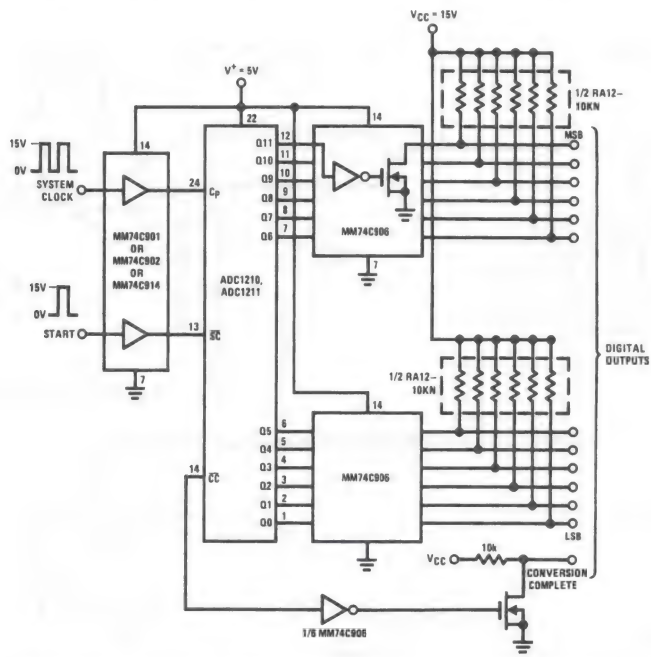
Several recommended operating configurations are shown in *Figure 5*.

Applications Information (Continued)



TL/H/5677–11

FIGURE 3. Interfacing an ADC1210, ADC1211 Running on $V^+ > V_{CC}$. Example: $V^+ = 10.24V$, System $V_{CC} = 5V$



TL/H/5677–12

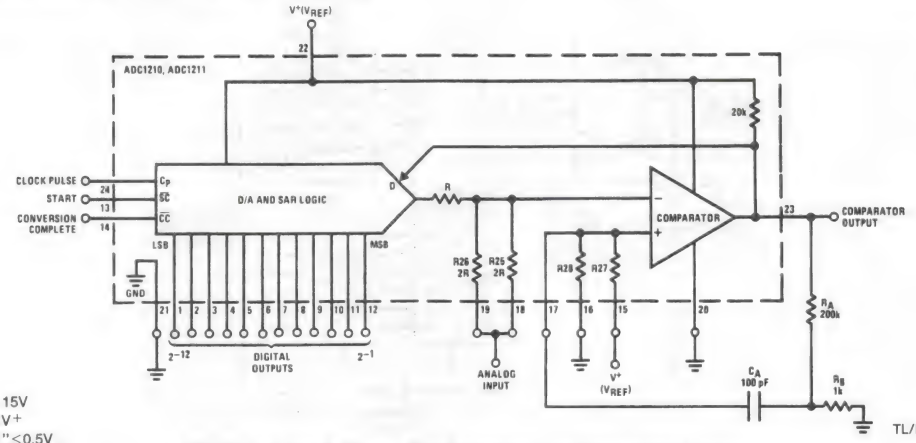
FIGURE 4. Interfacing an ADC1210, ADC1211 Running on $V^+ < V_{CC}$. Example: $V^+ = 5V$, $V_{CC} = 15V$

2.5 Offset and Full Scale Adjust

A variety of techniques may be employed to adjust Offset and Full Scale on the ADC1210, ADC1211. A straight-forward Full Scale Adjust is to incrementally vary V^+ (V_{REF}) to match the analog input voltage. A recommended technique is shown in *Figure 6*. An LM199 and low drift op amp(e.g., the LH0044) are used to provide the precision reference. The ADC1210, ADC1211 is put in the continuous convert mode by shorting pins 13 and 14. An analog voltage equal to V_{REF} minus 1½ LSB (10.23625V) is applied to pins 18 and 19, and R1 is adjusted until the LSB flickers equally between logic "1" and logic "0" (all other outputs must be stable logic "0"). Offset Null is accomplished by then applying an analog input voltage equal to ½ LSB at pins 18 and 19. R2 is adjusted until the LSB output flickers equally between logic "1" and logic "0" (all other bits are stable). In the circuit of *Figure 6*, the ADC1210, ADC1211 is configured for Complementary Binary logic and the values shown are for $V^+ = 10.240V$, $V_{FS} = 10.2375V$, LSB = 2.5 mV.

An alternate technique is shown in *Figure 7*. In this instance, an LH0071 is used to provide the reference voltage. An analog input voltage equal to V_{REF} minus 1½ LSB (10.23625V) is applied to pins 18 and 19.

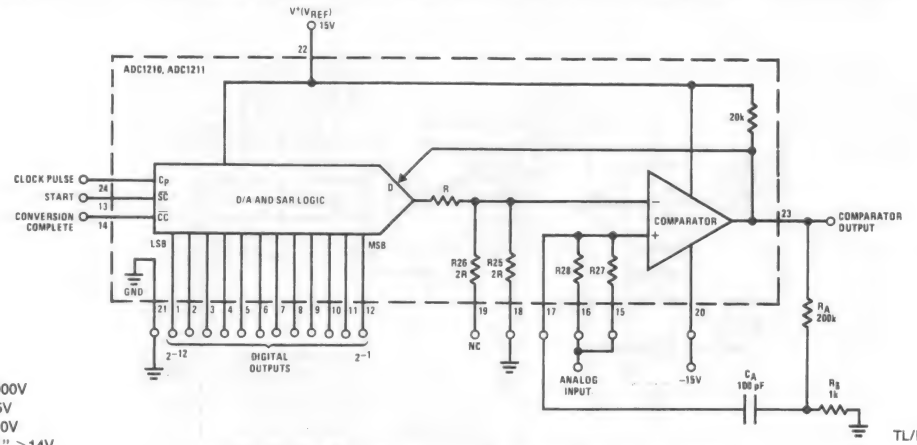
Applications Information (Continued)



TL/H/5677–13

$5V \leq V^+ \leq 15V$
$0V \leq V_{IN} \leq V^+$
Logical "1" $\leq 0.5V$
Logical "0" $\cong V^+$

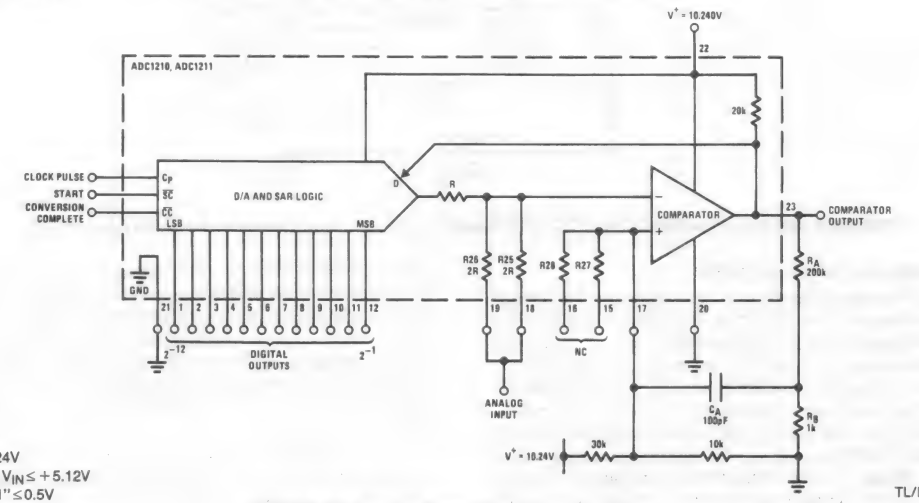
FIGURE 5a. Single Supply Configuration, Complementary Logic



TL/H/5677–14

$V^+ = 15.000V$
$V^- = -15V$
$0 \leq V_{IN} \leq 10V$
Logical "1" $\geq 14V$
Logical "0" $\leq 0.5V$

FIGURE 5b. High Voltage CMOS Compatible, 0V to 10V Input

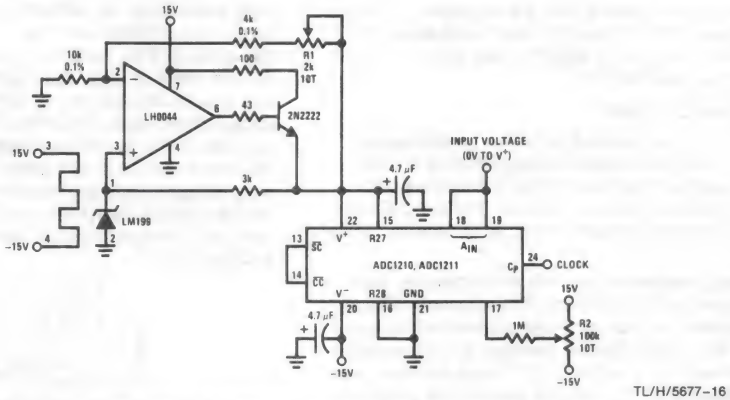


TL/H/5677–15

$V^+ = 10.24V$
$-5.12V \leq V_{IN} \leq +5.12V$
Logical "1" $\leq 0.5V$
Logical "0" $\cong 10V$

FIGURE 5c. Bipolar Input, Complementary Logic

Applications Information (Continued)



TL/H/5677–16

FIGURE 6. Offset and Full Scale Adjustment for Complementary Binary

R1 is adjusted until the LSB output flickers equally between logic "1" and logic "0" (all other outputs must be a stable logic "0"). For Offset Null, an analog voltage equal to 1/2 LSB (1.25 mV) is then applied to pins 18 and 19, and R2, is adjusted until the LSB output flickers equally between logic "1" and "0".

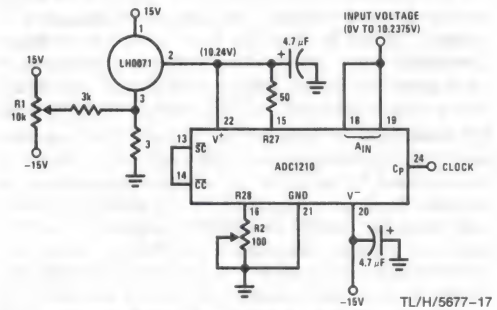


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FIGURE 7. Offset and Full-Scale Adjustment Technique Using LH0071

In both techniques shown, adjusting the Full-Scale first and then Offset minimizes adjustment interaction. At least one iteration is recommended as a self-check.

2.6 START PULSE CONSIDERATIONS

To assure reliable conversion accuracy, the $\overline{\text{START}}$ ($\overline{\text{SC}}$) pulse applied to pin 13 of the ADC1210 should be synchronized to the conversion clock. One simple way to do that is the circuit shown in *Figure 8*. Note that once a conversion cycle is initiated, the $\overline{\text{START}}$ signal cannot effect the conversion operation until it is completed.

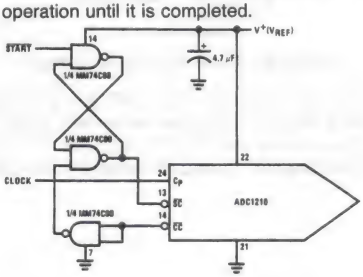

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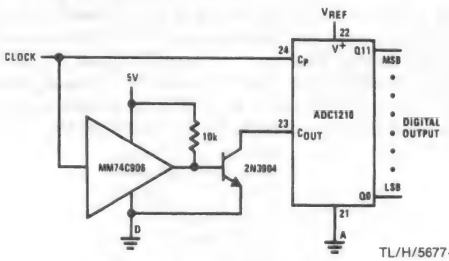
FIGURE 8. Synchronizing the $\overline{\text{START}}$ Pulse

The circuit insures that in no case can the ADC1210 make an error in the Most Significant Bit (MSB) decision. Without the circuit, it is possible for energy from the trailing edge of an asynchronous $\overline{\text{START}}$ pulse to be coupled into the ADC1210's comparator. If the analog input is near half-scale, the charge injected can force an error in the MSB decision. The circuit allows one clock period for this energy to dissipate before the decision is recorded.

2.7 ADC1210 CONVERSION AT 26 μs

The ADC1210 can run at 500 kHz clock frequency, or 12-bit conversion time of 26 μs (*Figure 9*). The comparator output is clamped low until the successive approximation register (SAR) is ready to strobe in the data at the rising edge of the conversion clock. Comparator oscillation is suppressed and kept from influencing the conversion decisions, eliminating the need for the AC hysteresis circuit above clock frequency of 65 kHz that is recommended.



TL/H/5677–18

FIGURE 9. Conversion at 26 μs

A complementary phased clock is required. The positive phase is used to clock the converter SAR as is normally the case. The same signal is buffered and inverted by the transistor. The open collector is wire-ORed to the output of the comparator. During the first half of the clock cycle (50% duty cycle), the comparator output is clamped and disabled, though its internal operation is still in normal working order. The last half cycle of the clock unclamps the comparator output. Thus, the output is permitted to slew to the final logic state just before the decision is logged into the SAR. The MM74C906 buffer (or with two inverting buffers) provides adequate propogation delay such that the comparator output data is held long enough to resolve any internal logic setup time requirements.

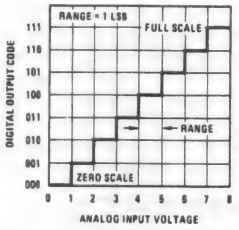
Applications Information (Continued)

The 500 kHz clock implies that the absolute minimum amount of time for the comparator output is *unclamped* is 1 μs. Therefore, if the clock is not 50% duty cycle, this 1 μs requirement must be observed.

3.0 DEFINITION OF TERMS

Resolution: The Resolution of an A/D is an expression of the smallest change in input which will increment (or decrement) the output from one code to the next adjacent code. It is defined in number of bits, or 1 part in 2^n. The ADC1210 and ADC1211 have a resolution of 12 bits or 1 part in 4,096 (0.0244%).

Quantization Uncertainty: Quantization Uncertainty is a direct consequence of the resolution of the converter. All analog voltages within a given range are represented by a single digital output code. There is, therefore, an inherent conversion error even for a perfect A/D. As an example, the transfer characteristic of a perfect 3-bit A/D is shown in *Figure 10*.



TL/H/5677–21

FIGURE 10. Quantization Uncertainty of a Perfect 3-Bit A/D

As can be seen, all input voltages between 0V and 1V are represented by an output code of 000. All input voltages between 1V and 2V are represented by an output code of 001, etc. If the midpoint of the range is assumed to be the nominal value (e.g., 0.5V), there is an Uncertainty of ±1/2 LSB. It is common practice to offset the converter 1/2 LSB in order to reduce the Uncertainty to ±1/2 LSB is shown in *Figure 11*, rather than +1, −0 shown in *Figure 10*. Quantization Uncertainty can only be reduced by increasing Resolution. It is expressed as ±1/2 LSB or as an error percentage of full scale (±0.0122% FS for the ADC1210).

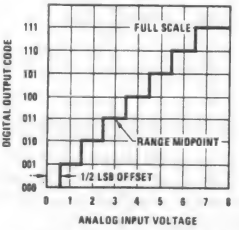


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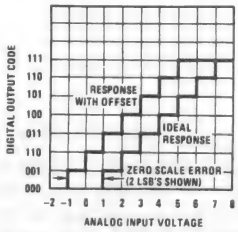
FIGURE 11. Transfer Characteristic Offset 1/2 LSB to Minimize Quantizing Uncertainty

Linearity Error: Linearity Error is the maximum deviation from a straight line passing through the end points of the A/D transfer characteristic. It is measured after calibrating Zero and Full Scale Error. Linearity is a performance characteristic intrinsic to the device and cannot be externally adjusted.

Zero Scale Error (or Offset): Zero Scale Error is a measure of the difference between the output of an ideal and the actual A/D for zero input voltage. As shown in *Figure 12*, the effect of Zero Scale Error is to shift the transfer characteristic to the right or left along the abscissa. Any voltage more negative than the LSB transition gives an output code of 000. In practice, therefore, the voltage at which the 000 to 001 transition takes place is ascertained, this input voltage's departure from the ideal value is defined as the Zero Scale Error (Offset) and is expressed as a percentage of FS. In the example of *Figure 12*, the offset is 2 LSB's or 0.286% of FS.



TL/H/5677–20

FIGURE 12. A/D Transfer Characteristic with Offset

The Zero Scale Error of the ADC1210, ADC1211 is caused primarily by offset voltage in the comparator. Because it is common practice to offset the A/D 1/2 LSB to minimize Quantization Error, the offsetting techniques described in the Applications Section may be used to null Zero Scale Error and accomplish the 1/2 LSB offset at the same time.

Full Scale Error (or Gain Error): Full Scale Error is a measure of the difference between the output of an ideal A/D converter and the actual A/D for an input voltage equal to full scale. As shown in *Figure 13*, the Full Scale Error effect is to rotate the transfer characteristic angularly about the origin. Any voltage more positive than the Full Scale transition gives an output code of 111. In practice, therefore, the voltage at which the transition from 111 to 110 occurs is ascertained. The input voltage's departure from the ideal value is defined as Full Scale Error and is expressed as a percentage of FS. In the example of *Figure 13*, Full Scale Error is 1 1/2 LSB's or 0.214% of FS.

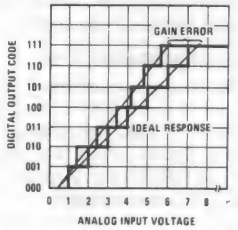


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FIGURE 13. Full Scale (Gain Error)

Full Scale Error of the ADC1210, ADC1211 is due primarily to mismatch in the R-2R ladder equivalent output impedance and input resistors R25, R26, R27, and R28. The gain error may be adjusted to zero as outlined in section 2.5.

Applications Information (Continued)

Monotonicity and Missing Codes: Monotonicity is a property of a D/A which requires an increasing or constant output voltage for an increasing digital input code. Monotonicity of a D/A converter does not, in itself, guarantee that an A/D built with that D/A will not have missing codes. However, the ADC1210 and ADC1211 are guaranteed to have no missing codes.

Conversion Time: The ADC1210, ADC1211 are successive approximation A/D converters requiring 13 clock intervals for a conversion to specified accuracy for the ADC1210 and 11 clocks for the ADC1211. There is a trade-off between accuracy and clock frequency due to settling time of the ladder and propagation delay through the comparator. By modifying the hysteresis network around the comparator, conversions with 10-bit accuracy can be made in 30 μs. Replace R_A, R_B and C_A in *Figure 5* with a 10 MΩ resistor between pin 23 (Comparator Output) and pin 17 (+ IN), and increase the clock rate to 366 kHz.

In order to prevent errors during conversion, the analog input voltage should not be allowed to change by more than $\pm 1/2$ LSB. This places a maximum slew rate of 12.5 μV/μs on the analog input voltage. The usual solution to this restriction is to place a Sample and Hold in front of the A/D. For additional application information, refer to application note AN245.

National Semiconductor

PRELIMINARY

ADC3511 3½-Digit Microprocessor Compatible A/D Converter
ADC3711 3¾-Digit Microprocessor Compatible A/D Converter

General Description

The ADC3511 and ADC3711 (MM74C937, MM74C938-1) monolithic A/D converter circuits are manufactured using standard complementary MOS (CMOS) technology. A pulse modulation analog-to-digital conversion technique is used and requires no external precision components. In addition, this technique allows the use of a reference voltage that is the same polarity as the input voltage.

One 5V (TTL) power supply is required. Operating with an isolated supply allows the conversion of positive as well as negative voltages. The sign of the input voltage is automatically determined and indicated on the sign pin. If the power supply is not isolated, only one polarity of voltage may be converted.

The conversion rate is set by an internal oscillator. The frequency of the oscillator can be set by an external RC network or the oscillator can be driven from an external frequency source. When using the external RC network, a square wave output is available.

The ADC3511 and ADC3711 have been designed to provide addressed BCD data and are intended for use with microprocessors and other digital systems. BCD digits are selected on demand via 2 Digit Select (D0, D1) inputs. Digit Select inputs are latched by a low-to-high transition on the Digit Latch Enable (DLE) input and will remain latched as long as DLE remains high. A start conversion input and a conversion complete output are included on both the ADC3511 and the ADC3711.

Features

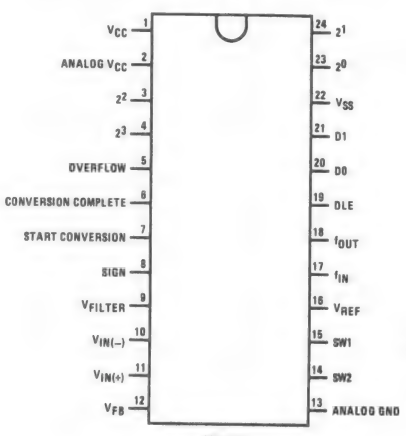
- Operates from single 5V supply
- ADC3511 converts 0 to ±1999 counts
- ADC3711 converts 0 to ±3999 counts
- Addressed BCD outputs
- No external precision components necessary
- Easily interfaced to microprocessors or other digital systems
- Medium speed—200 ms/conversion
- TTL compatible
- Internal clock set with RC network or driven externally
- Overflow indicated by hex "EEEE" output reading as well as an overflow output
- ADC3511 equivalent to MM74C937
- ADC3711 equivalent to MM74C938-1

Applications

- Low cost analog-to-digital converter
- Eliminate analog multiplexing by using remote A/D converters
- Convert analog transducers (temperature, pressure, displacement, etc.) to digital transducers

Connection Diagram

Dual-In-Line Package



TOP VIEW

Order Number ADC3511CCN or ADC3711CCN
NS Package N24A

TL/H/5678–1

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Voltage at Any Pin	$-0.3V$ to $V_{CC}+0.3V$
Operating Temperature Range (T_A)	$-40°C$ to $+85°C$
Package Dissipation at $T_A=25°C$	500 mW
Operating V_{CC} Range	4.5V to 6.0V
Absolute Maximum V_{CC}	6.5V
Storage Temperature Range	$-65°C$ to $+150°C$
Lead Temp. (Soldering, 10 seconds)	260°C
ESD Susceptibility (Note 5)	TBD V

DC Electrical Characteristics ADC3511CC, ADC3711CC

$4.75V \le V_{CC} \le 5.25V$, $-40°C \le T_A \le +85°C$, unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ (Note 2)	Max	Units
$V_{IN(1)}$	Logical "1" Input Voltage (Except f_{IN})		$V_{CC}-1.5$			V
$V_{IN(0)}$	Logical "0" Input Voltage (Except f_{IN})				1.5	V
$V_{IN(1)}$	Logical "1" Input Voltage (f_{IN})		$V_{CC}-0.6$			V
$V_{IN(0)}$	Logical "0" Input Voltage (f_{IN})				0.6	V
$V_{OUT(1)}$	Logical "1" Output Voltage (Except 2^0, 2^1, 2^2, 2^3)	$I_O=360\ \mu A$	$V_{CC}-0.4$			V
$V_{OUT(1)}$	Logical "1" Output Voltage (2^0, 2^1, 2^2, 2^3)	$I_O=360\ \mu A$	$V_{CC}-1.0$			V
$V_{OUT(0)}$	Logical "0" Output Voltage	$I_O=1.6$ mA			0.4	V
$I_{IN(1)}$	Logical "1" Input Current (SC, DLE, D0, D1)	$V_{IN}=V_{CC}$		0.005	1.0	μA
$I_{IN(0)}$	Logical "0" Input Current (SC, DLE, D0, D1)	$V_{IN}=0V$	-1.0	-0.005		μA
I_{CC}	Supply Current	All Outputs Open		0.5	5.0	mA

AC Electrical Characteristics ADC3511CC, ADC3711CC

$V_{CC}=5V$; $T_A=25°C$, $C_L=50$ pF; $t_r=t_f=20$ ns; unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ (Note 2)	Max	Units
f_{OSC}	Oscillator Frequency			0.6/RC		Hz
f_{IN}	Clock Frequency		100		640	kHz
f_{CONV}	Conversion Rate	ADC3511CC ADC3711CC		$f_{IN}/64{,}512$ $f_{IN}/129{,}024$		conversions/sec conversions/sec
t_{SCPW}	Start Conversion Pulse Width		200		DC	ns
t_{pd0}, t_{pd1}	Propagation Delay D0, D1, to 2^0, 2^1, 2^2, 2^3	DLE = 0V		2.0	5.0	μs
t_{pd0}, t_{pd1}	Propagation Delay DLE to 2^0, 2^1, 2^2, 2^3			2.0	5.0	μs
t_{SET-UP}	Set-Up Time D0, D1, to DLE	$t_{HOLD}=0$ ns		100	200	ns
t_{PWDLE}	Minimum Pulse Width Digit Latch Enable (Low)			100	200	ns

Converter Characteristics

ADC3511CC, ADC3711CC $4.75 \le V_{CC} \le 5.25V$; $-40°C \le T_A \le +85°C$, $f_C = 5$ conv./sec (ADC3511CC); 2.5 conv./sec (ADC3711CC); unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ (Note 2)	Max	Units
	Non-Linearity	$V_{IN} = 0-2V$ Full Scale $V_{IN} = 0-200$ mV Full Scale	−0.05	±0.025	+0.05	% of Full-Scale (Note 3)
	Quantization Error		−1		+0	Counts
	Offset Error	$V_{IN} = 0V$	−0.5	+1.0	+3.0	mV (Note 4)
	Rollover Error		−0		+0	Counts
$V_{IN}+$, $V_{IN}-$	Analog Input Current	$T_A = 25°C$	−5	±1	+5	nA

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All typicals are given for $T_A = 25°C$.

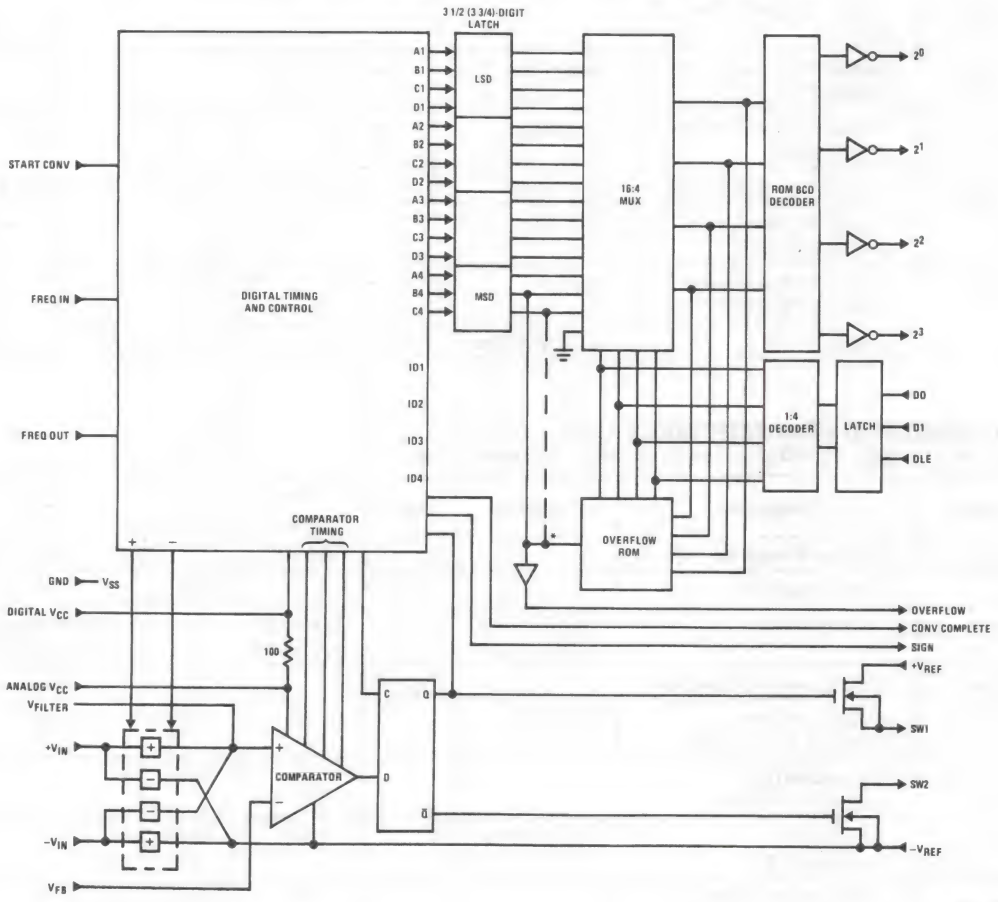
Note 3: For the ADC3511CC: full-scale = 1999 counts; therefore 0.025% of full-scale = ½ count and 0.05% of full-scale = 1 count. For the ADC3711CC: full-scale = 3999 counts; therefore 0.025% of full-scale = 1 count and 0.05% of full-scale = 2 counts.

Note 4: For full-scale = 2.000V: 1 mV = 1 count for the ADC3511CC; 1 mV = 2 counts for the ADC3711CC.

Note 5: Human body model, 100 pF discharged through a 1.5Ω resistor.

Block Diagram

ADC3511 3 ½-Digit A/D (*ADC3711 3 ¾-Digit A/D)



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Applications Information

THEORY OF OPERATION

A schematic for the analog loop is shown in *Figure 1.* The output of SW1 is either at V_{REF} or zero volts, depending on the state of the D flip-flop. If Q is at a high level, $V_{OUT} = V_{REF}$ and if Q is at a low level $V_{OUT} = 0V$. This voltage is then applied to the low pass filter comprised of R1 and C1. The output of this filter, V_{FB}, is connected to the negative input of the comparator, where it is compared to the analog input voltage, V_{IN}. The output of the comparator is connected to the D input of the D flip-flop. Information is then transferred from the D input to the Q and $\overline{Q}$ outputs on the positive edge of clock. This loop forms an oscillator whose duty cycle is precisely related to the analog input voltage, V_{IN}.

An example will demonstrate this relationship. Assume the input voltage is equal to 0.500V. If the Q output of the D flip-flop is high then V_{OUT} will equal V_{REF} (2.000V) and V_{FB} will charge toward 2V with a time constant equal to R1C1. At some time V_{FB} will exceed 0.500V and the comparator output will switch to 0V. At the next clock rising edge the Q output of the D flip-flop will switch to ground, causing V_{OUT} to switch to 0V. At this time, V_{FB} will start discharging toward 0V with a time constant R1C1. When V_{FB} is less than 0.5V the comparator output will switch high. On the rising edge of the next clock the Q output of the D flip-flop will switch high and the process will repeat. There exists at the output of SW1 a square wave pulse train with positive amplitude V_{REF} and negative amplitude 0V.

The DC value of this pulse train is:

$$V_{OUT} = V_{REF} \frac{t_{ON}}{t_{ON} + t_{OFF}} = V_{REF} \text{ (duty cycle)}$$

The lowpass filter will pass the DC value and then:

$$V_{FB} = V_{REF} \text{ (duty cycle)}$$

Since the closed loop system will always force V_{FB} to equal V_{IN}, we can then say that:

$$V_{IN} = V_{FB} = V_{REF} \text{ (duty cycle)}$$

or

$$\frac{V_{IN}}{V_{REF}} = \text{(duty cycle)}$$

The duty cycle is logically ANDed with the input frequency f_{IN}. The resultant frequency f equals:

$$f = (\text{duty cycle}) \times (f_{IN})$$

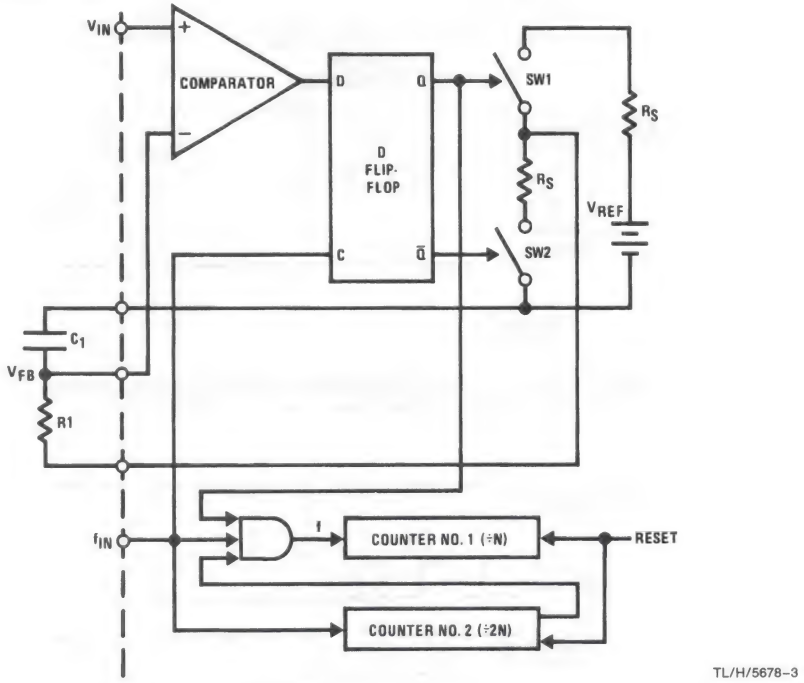
Frequency f is accumulated by counter no. 1 for a time determined by counter no. 2. The count contained in counter no. 1 is then:

$$(\text{count}) = \frac{f}{(f_{IN})/N} = \frac{(\text{duty cycle}) \times (f_{IN})}{(f_{IN})/N}$$

$$= \frac{V_{IN}}{V_{REF}} \times N$$

For the ADC3511 N = 2000.

For the ADC3711 N = 4000.



$V_{IN} = V_{FB} = V_{REF} \times (\text{duty cycle})$

$f = (\text{duty cycle}) \times f_{IN}$

$$\text{Count in counter no. 1} = \frac{f}{(f_{IN})/N} = \frac{(\text{duty cycle}) \times f_{IN}}{(f_{IN})/N} = \frac{V_{IN}}{V_{REF}} \times N$$

FIGURE 1. Analog Loop Schematic Pulse Modulation A/D Converter

Applications Information (Continued)

GENERAL INFORMATION

The timing diagram, shown in *Figure 2,* gives operation for the free running mode. Free running operation is obtained by connecting the Start Conversion input to logic "1" (V_{CC}). In this mode the analog input is continuously converted and the digit latches are updated at a rate equal to 64,512 × $1/f_{IN}$ for the ADC3511, or 129,024 for the ADC3711.

The rising edge of the Conversion Complete output indicates that new information has been transferred from the internal counter to the digit latches. This information will remain in the digit latches until the next low-to-high transition of the Conversion Complete output. A logic "1" will be maintained on the Conversion Complete output for a time equal to 64×$1/f_{IN}$ on the ADC3511, or 128×$1/f_{IN}$ on the ADC3711.

Figure 3 gives the operation using the Start Conversion input. It is important to note that the Start Conversion input and Conversion Complete output do not influence the actual analog-to-digital conversion in any way. Internally the ADC3511 and ADC3711 are always continuously converting the analog voltage present at their inputs. The Start Conversion input is used to control the transfer of information from the internal counter to the digit latches.

An RS latch on the Start Conversion input allows a broad range of input pulse widths to be used on this signal. As shown in *Figure 3,* the Conversion Complete output goes to a logic "0" on the rising edge of the Start Conversion pulse and goes to a logic "1" some time later when the new conversion is transferred from the internal counter to the display latch. Since the Start Conversion pulse can occur at any time during the conversion cycle, the amount of time from Start Conversion to Conversion Complete will vary. The maximum time is 64,512 × $1/f_{IN}$ (129,024 × $1/f_{IN}$ for the ADC3711) and the minimum time is 256 × $1/f_{IN}$ (512 × $1/f_{IN}$) for the ADC3711).

SYSTEM DESIGN CONSIDERATIONS

The ADC3511 and ADC3711 have reduced the problem of high resolution, high accuracy analog-to-digital conversion to nearly the level of simplicity, economy, and compactness usually associated with digital logic circuitry. However, they are truly high precision analog devices, and require the same kind of design considerations given to all analog circuits. While great care has been taken in the design of the ADC3511 and ADC3711 to make their application as easy as possible, in order to utilize them to their full performance potential, good grounding, power supply distribution, decoupling, and regulation techniques should be exercised.

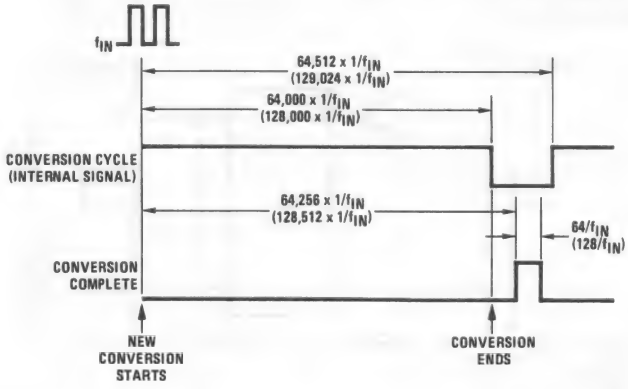


FIGURE 2. Conversion Cycle Timing Diagram for Free Running Operation (Times Shown in Parentheses are for the ADC3711)

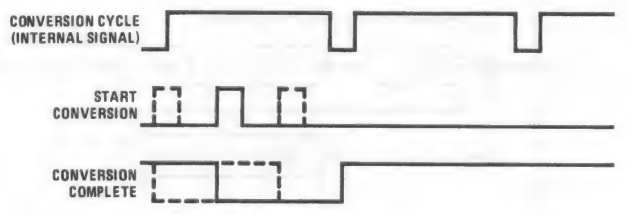


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FIGURE 3. Conversion Cycle Timing Diagram Operating with Start Conversion Input

Truth Table

DIGIT SELECT INPUTS			SELECTED DIGIT
DLE	D1	D0	
L	L	L	Digit 0 (LSD)
L	L	H	Digit 1
L	H	L	Digit 2
L	H	H	Digit 3 (MSD)
H	X	X	Unchanged

L = low logic level

H = high logic level

X = irrelevant logic level

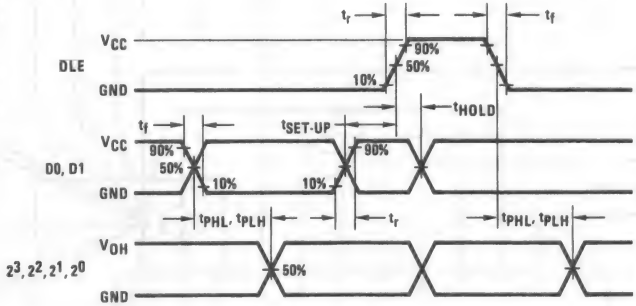
The value of the Selected Digit is presented at the 2^3, 2^2, 2^1 and 2^0 outputs in BCD format.

Note 1: If the value of a digit changes while it is selected, that change *will* be reflected at the outputs.

Note 2: An overflow condition will be indicated by a high level on the OVERFLOW output (pin 5) and E16 in all digits.

Note 3: The sign of the input voltage, when these devices are operated in the bipolar mode, is indicated by the SIGN output (pin 8). A high level indicates a positive voltage, a low level a negative.

Timing Diagrams



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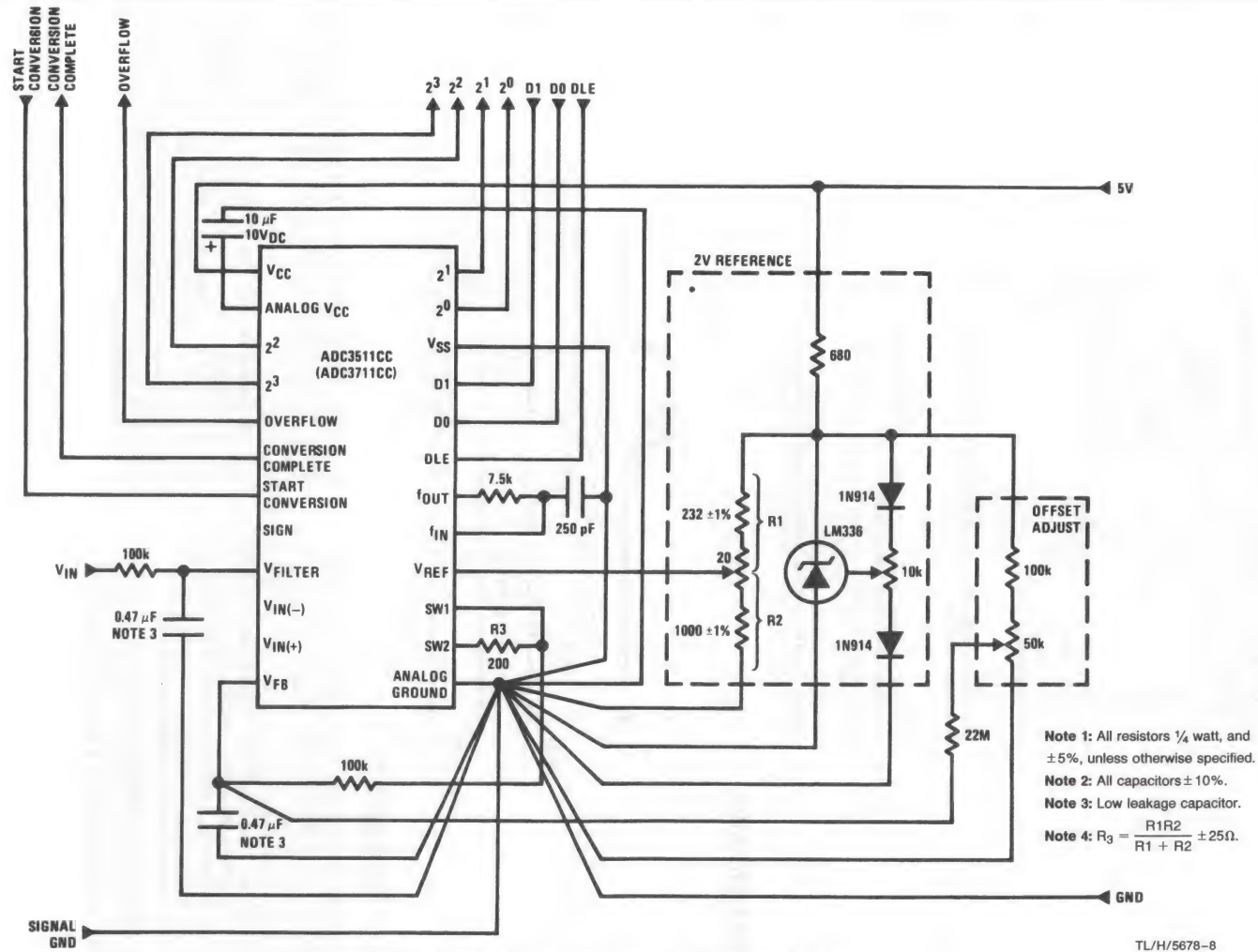
Typical Applications

Figure 4 shows the ADC3511 and ADC3711 connected to convert 0 to +2.000 volts full scale operating from a non-isolated power supply. (Note that the ADC3511 converts 0 to +1999 counts full scale, while the ADC3711 converts 0 to +3999 counts full scale.) In this configuration the SIGN output (pin 8) should be ignored. Higher voltages can, of course, be converted by placing fixed dividers in the inputs, while lower voltages can be converted by placing fixed dividers in the feedback loop, as shown in *Figure 6.*

Figures 5 and *6* show systems operating with isolated supplies that will convert both polarities of inputs. 60 Hz common-mode noise can become a problem in these configurations, so shielded transformers have been shown in the figures. The necessity for, and the type of shielding needed depends on the performance requirements, and the actual applications.

The filter capacitors connected to V_{FB} (pin 12) and V_{FILTER} (pin 11) should be of a low leakage variety. In the examples shown every 1.0 nA of leakage will cause approximately 0.1 mV error ($1.0 \times 10^{-9}A \times 100\ k\Omega = 0.1$ mV). If the currents in both capacitors are exactly equal however, little error will result since the source impedances driving both capacitors are approximately matched.

Typical Applications (Continued)



Note 1: All resistors ¼ watt, and ±5%, unless otherwise specified.

Note 2: All capacitors ± 10%.

Note 3: Low leakage capacitor.

Note 4: $R_3 = \frac{R1R2}{R1 + R2} \pm 25\Omega$.

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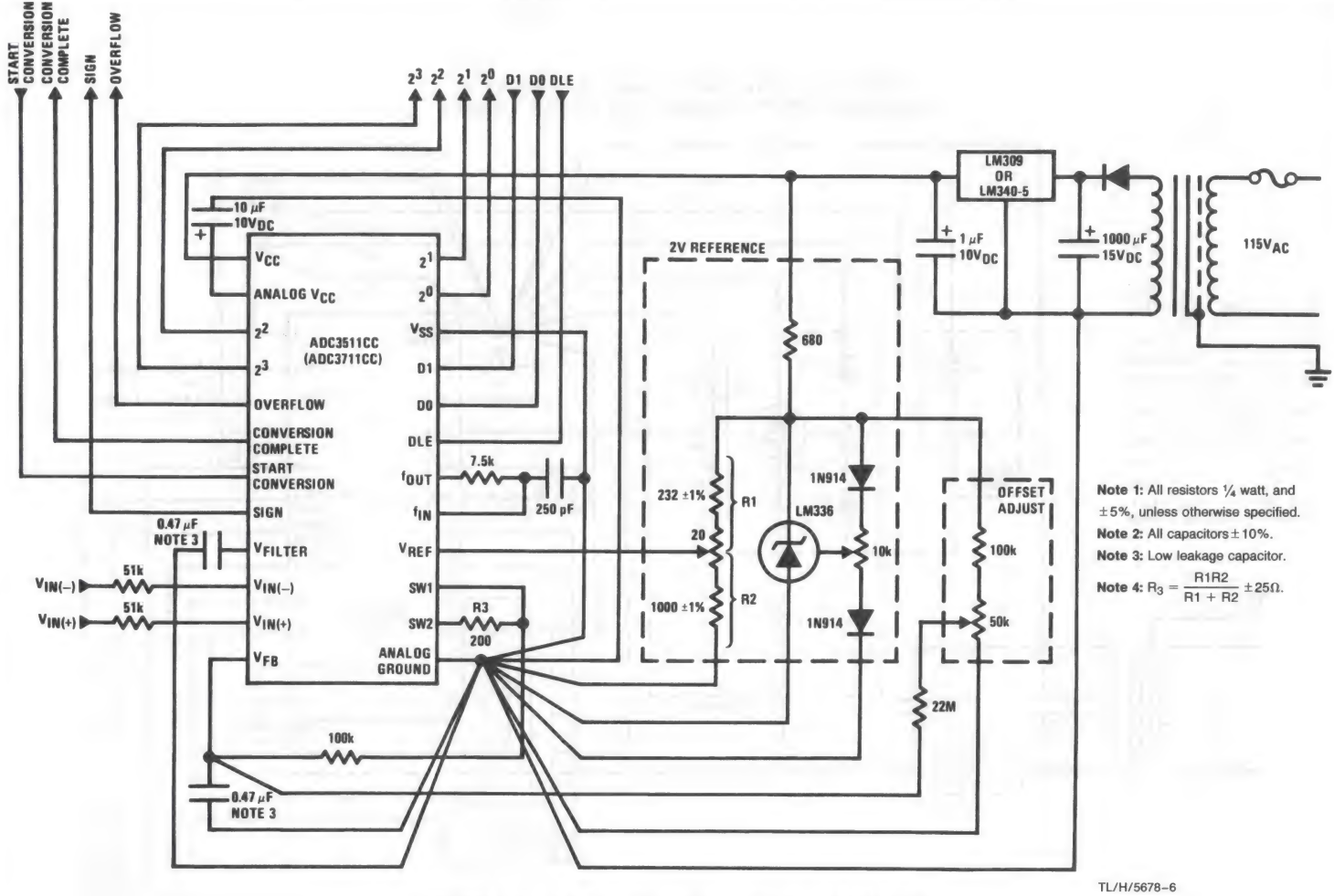
FIGURE 4. 3 ½-Digit A/D; + 1999 Counts, + 2.000 Volts Full Scale
(3 ¾-Digit A/D; + 3999 Counts, + 2.000 Volts Full Scale)



Note 1: All resistors ¼ watt, and ±5%, unless otherwise specified.
Note 2: All capacitors ±10%.
Note 3: Low leakage capacitor.
Note 4: $R_3 = \frac{R1R2}{R1 + R2} \pm 25\Omega$.

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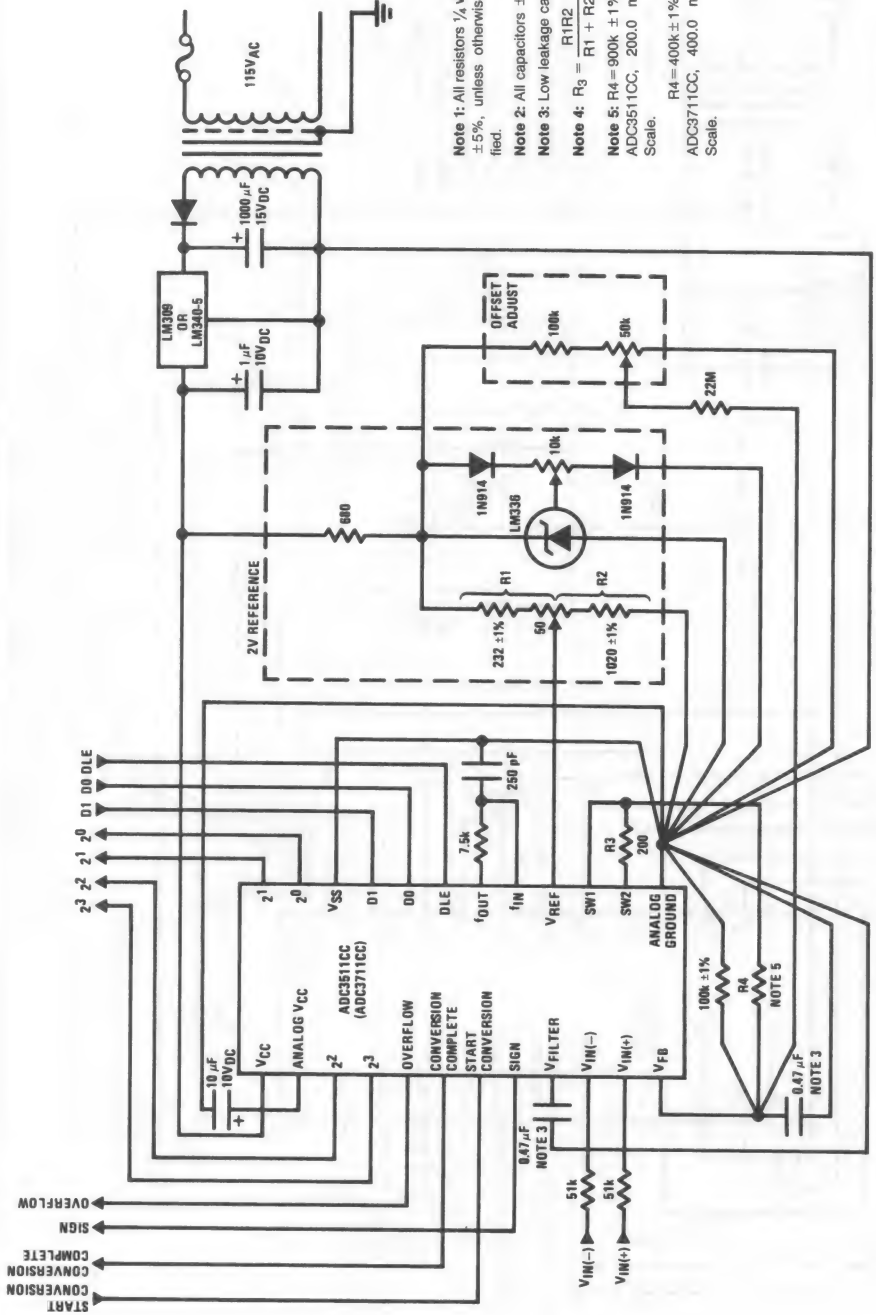
FIGURE 5. 3 ½-Digit A/D; ±1999 Counts, ±2.000 Volts Full Scale
(3 ¾-Digit A/D; ±3999 Counts, ±2.000 Volts Full Scale)

Typical Applications (Continued)



Note 1: All resistors ¼ watt, and ±5%, unless otherwise specified.

Note 2: All capacitors ±10%

Note 3: Low leakage capacitor.

Note 4: $R_3 = \frac{R1R2}{R1 + R2} \pm 50\Omega$

Note 5: R4 = 900k ±1% for the ADC3511CC, 200.0 mV Full-Scale.

R4 = 400k ±1% for the ADC3711CC, 400.0 mV Full-Scale.

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FIGURE 6. 3 ½-Digit A/D; ±1999 Counts, ±200.0 mV Full Scale (3 ¾-Digit A/D; ±3999 Counts, ±400.0 mV Full-Scale)

National Semiconductor

ADD3501 3½ Digit DVM with Multiplexed 7-Segment Output

General Description

The ADD3501 monolithic DVM circuit is manufactured using standard complementary MOS (CMOS) technology. A pulse modulation analog-to-digital conversion technique is used and requires no external precision components. In addition, this technique allows the use of a reference voltage that is the same polarity as the input voltage.

One 5V (TTL) power supply is required. Operating with an isolated supply allows the conversion of positive as well as negative voltages. The sign of the input voltage is automatically determined and output on the sign pin. If the power supply is not isolated, only one polarity of voltage may be converted.

The conversion rate is set by an internal oscillator. The frequency of the oscillator can be set by an external RC network or the oscillator can be driven from an external frequency source. When using the external RC network, a square wave output is available. It is important to note that great care has been taken to synchronize digit multiplexing with the A/D conversion timing to eliminate noise due to power supply transients.

The ADD3501 has been designed to drive 7-segment multiplexed LED displays directly with the aid of external digit buffers and segment resistors. Under condition of overrange, the overflow output will go high and the display will read +OFL or −OFL, depending on whether the input voltage is positive or negative. In addition to this, the most significant digit is blanked when zero.

A start conversion input and a conversion complete output are included on all 4 versions of this product.

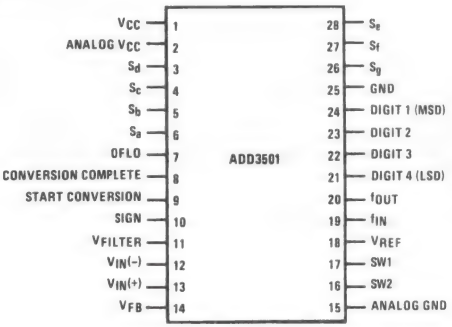
Features

- Operates from single 5V supply
- Converts 0V to ±1.999V
- Multiplexed 7-segment
- Drives segments directly
- No external precision component necessary
- Accuracy specified over temperature
- Medium speed - 200ms/conversion
- Internal clock set with RC network or driven externally
- Overrange Indicated by +OFL or −OFL display reading and OFLO output
- Analog inputs in applications shown can withstand ±200 Volts
- ADD3501 equivalent to MM74C935

Applications

- Low cost digital power supply readouts
- Low cost digital multimeters
- Low cost digital panel meters
- Eliminate analog multiplexing by using remote A/D converters
- Convert analog transducers (temperature, pressure, displacement, etc.) to digital transducers

Connection Diagram



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Order Number ADD3501CCN
See NS Package Number N28B

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Voltage at Any Pin	−0.3V to V_{CC} +0.3V
Operating Temperature Range (T_A)	−40°C to +85°C
ESD Susceptibility (Note 3)	TBDV
Package Dissipation at T_A=25°C derate at $\theta_{JA(MAX)}$=125°C/Watt above T_A=25°C	800 mW
Operating V_{CC} Range	4.5V to 6.0V
Absolute Maximum V_{CC}	6.5V
Lead Temp. (Soldering, 10 seconds)	260°C
Storage Temperature Range	−65°C to +150°C

Electrical Characteristics ADD3501

4.75V ≤ V_{CC} ≤ 5.25V, −40°C ≤ T_A ≤ +85°C, unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ(2)	Max	Units
$V_{IN(1)}$	Logical "1" Input Voltage		V_{CC}−1.5			V
$V_{IN(0)}$	Logical "0" Input Voltage				1.5	V
$V_{OUT(0)}$	Logical "0" Output Voltage (All Digital Outputs except Digit Outputs)	I_O=1.1 mA			0.4	V
$V_{OUT(0)}$	Logical "0" Output Voltage (Digit Outputs)	I_O=0.7 mA			0.4	V
$V_{OUT(1)}$	Logical "1" Output Voltage (All Segment Outputs)	I_O=50 mA@T_J=25°C V_{CC}=5V I_O=30 mA@T_J=100°C	V_{CC}−1.6 V_{CC}−1.6	V_{CC}−1.3 V_{CC}−1.3		V V
$V_{OUT(1)}$	Logical "1" Output Voltage (All Digital Outputs except Segment Outputs)	I_O=500μA (Digit Outputs) I_O=360μA (Conv. Complete, +/−, Oflo Outputs)	V_{CC}−0.4			V
I_{SOURCE}	Output Source Current (Digit Outputs)	V_{OUT}=1.0V	2.0			mA
$I_{IN(1)}$	Logical "1" Input Current (Start Conversion)	V_{IN}=1.5V			1.0	μA
$I_{IN(0)}$	Logical "0" Input Current (Start Conversion)	V_{IN}=0V	−1.0			μA
I_{CC}	Supply Current	Segments and Digits Open		0.5	10	mA
f_{OSC}	Oscillator Frequency			0.6/RC		kHz
f_{IN}	Clock Frequency		100		640	kHz
f_C	Conversion Rate			f_{IN}/64,512		conv./sec
f_{MUX}	Digit Mux Rate			f_{IN}/256		Hz
t_{BLANK}	Inter Digit Blanking Time			1/(32f_{MUX})		sec
t_{SCPW}	Start Conversion Pulse Width		200		DC	ns

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All typicals given for T_A = 25°C.

Note 3: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

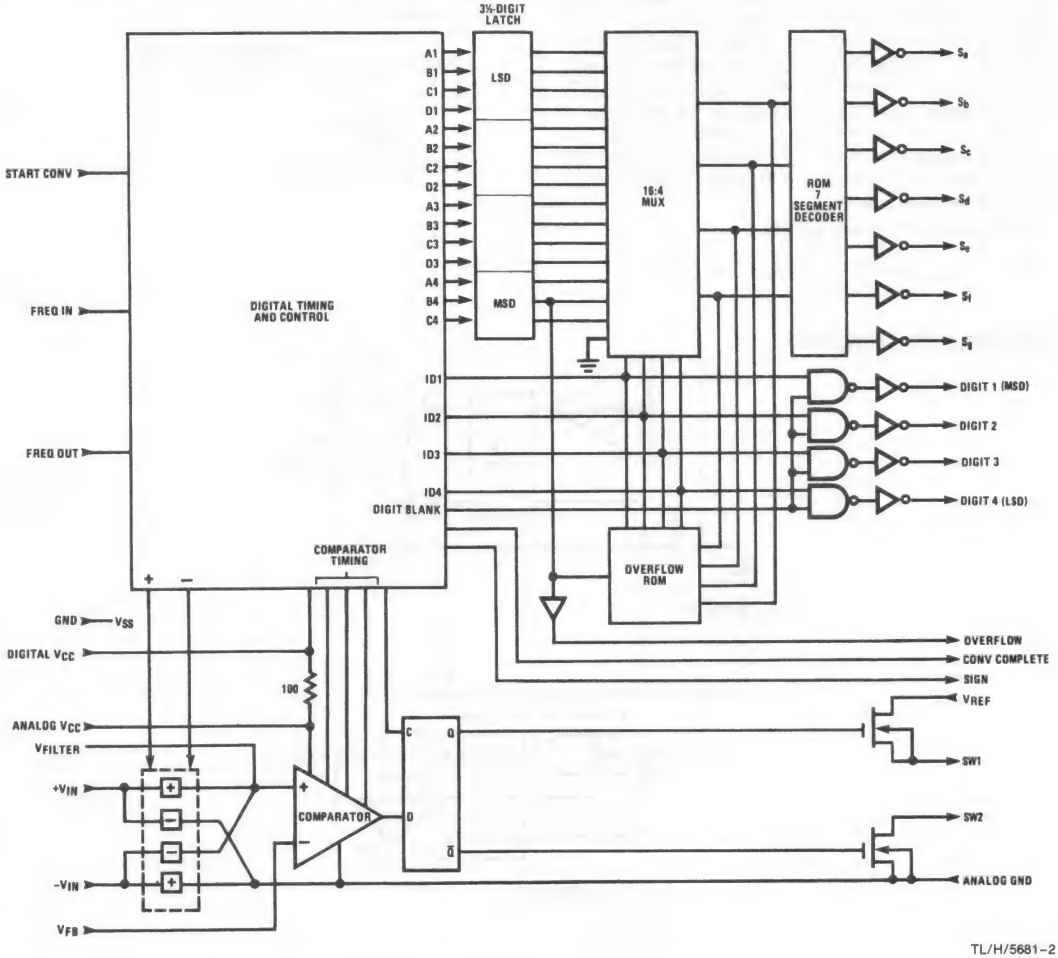
Electrical Characteristics ADD3501

t_C = 5 conversions/second, 0°C ≤ T_A ≤ 70°C, unless otherwise specified.

Parameter	Conditions	Min	Typ	Max	Units
Non-Linearity	V_{IN} = 0 − 2V Full Scale V_{IN} = 0 − 200mV Full Scale	−0.05 −0.05	±0.025 ±0.025	+0.05 +0.05	% of full scale
Quantization Error		−1		+0	counts
Offset Error, V_{IN} = 0V		−0.5	+1.5	+3	mV
Rollover Error		−0		+0	counts
Analog Input Current (V_{IN}+, V_{IN}−)	T_A = 25°C	−5	±0.5	+5	nA

Block Diagram

ADD3501 3½-Digit DVM Block Diagram



TL/H/5681–2

Theory of Operation

A schematic for the analog loop is shown in *Figure 1.* The output of SW1 is either at V_{REF} or zero volts, depending on the state of the D flip-flop. If Q is at a high level $V_{OUT} = V_{REF}$ and if Q is at a low level $V_{OUT} = 0V$. This voltage is then applied to the low pass filter comprised of R1 and C1. The output of this filter, V_{FB}, is connected to the negative input of the comparator, where it is compared to the analog input voltage, V_{IN}. The output of the comparator is connected to the D input of the D flip-flop. Information is then transferred from the D input to the Q and Q outputs on the positive edge of clock. This loop forms an oscillator whose duty cycle is precisely related to the analog input voltage, V_{IN}.

An example will demonstrate this relationship. Assume the input voltage is equal to 0.500V. If the Q output of the D flip-flop is high then V_{OUT} will equal V_{REF} (2.000V) and V_{FB} will charge toward 2V with a time constant equal to R_1C_1. At some time V_{FB} will exceed 0.500V and the comparator output will switch to 0V. At the next clock rising edge the Q output of the D flip-flop will switch to ground, causing V_{OUT} to switch to 0V. At this time V_{FB} will start discharging toward 0V with a time constant R_1C_1. When V_{FB} is less than 0.5V the comparator output will switch high. On the rising edge of the next clock the Q output of the D flip-flop will switch high and the process will repeat. There exists at the output of SW1 a square wave pulse train with positive amplitude V_{REF} and negative amplitude 0V.

The DC value of this pulse train is:

$$V_{OUT} = V_{REF}\left(\frac{T_{ON}}{T_{ON} + T_{OFF}}\right) = V_{REF}(\text{duty cycle})$$

The lowpass filter will pass the DC value and then:

$$V_{FB} = V_{REF}(\text{duty cycle})$$

Since the closed loop system will always force V_{FB} to equal V_{IN}, we can then say that:

$$V_{IN} = V_{FB} = V_{REF}(\text{duty cycle})$$

or

$$\frac{V_{IN}}{V_{REF}} = (\text{duty cycle})$$

The duty cycle is logically ANDed with the input frequency f_{IN}. The resultant frequency f equals:

$$f = (\text{duty cycle}) \times (\text{clock})$$

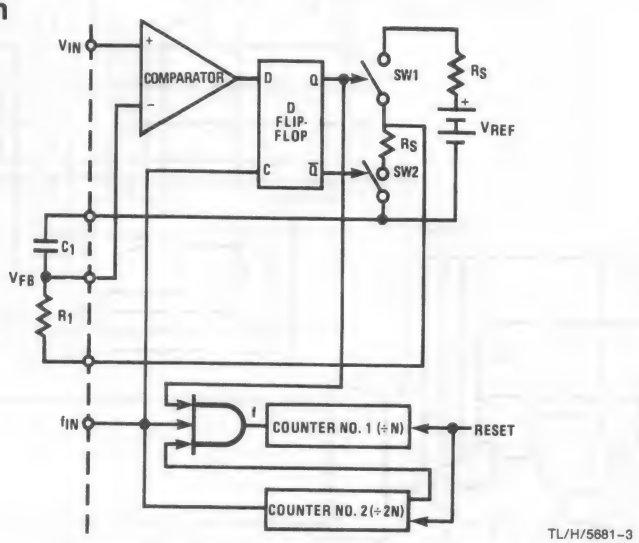
Frequency f is accumulated by counter no. 1 for a time determined by counter no. 2. The count contained in counter no. 1 is then:

$$(\text{count}) = \frac{f}{(\text{clock})/N} = \frac{(\text{duty cycle}) \times (\text{clock})}{(\text{clock})/N}$$

$$= \frac{V_{IN}}{V_{REF}} \times N$$

For the ADD3501, N = 2000.

Schematic Diagram



$V_{IN} = V_{FB} = V_{REF} \times (\text{duty cycle})$

$f = (\text{duty cycle}) \times f_{IN}$

$$\text{Count in Counter No. 1} = \frac{f}{f_{IN}/N} = \frac{(\text{duty cycle}) \times f_{IN}}{f_{IN}/N} = \frac{V_{IN}}{V_{REF}} \times N$$

Figure 1. Analog Loop Schematic
Pulse Modulation A/D Converter

General Information

The timing diagram, shown in *Figure 2*, gives operation for the free running mode. Free running operation is obtained by connecting the Start Conversion input to logic "1" (V_{CC}). In this mode the analog input is continuously converted and the display is updated at a rate equal to $64{,}512 \times 1/f_{IN}$.

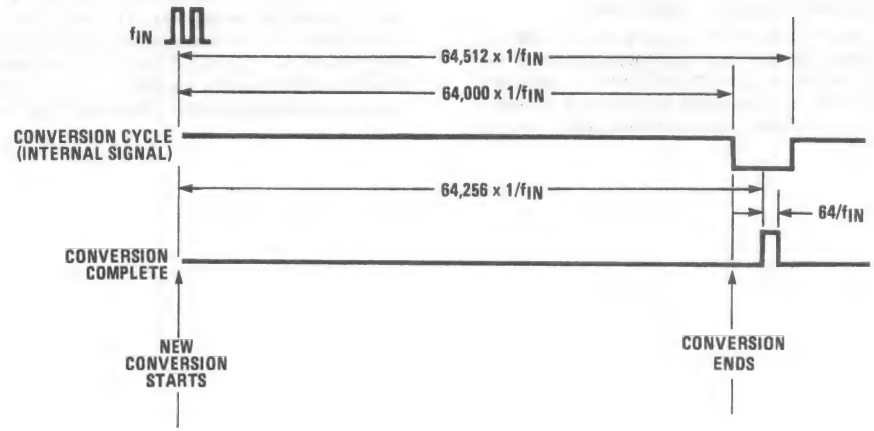
The rising edge of the Conversion Complete output indicates that new information has been transferred from the internal counter to the display latch. This information will remain in the display latch until the next low-to-high transition of the Conversion Complete output. A logic "1" will be maintained on the Conversion Complete output for a time equal to $64 \times 1/f_{IN}$.

Figure 3 gives the operation using the Start Conversion input. It is important to note that the Start Conversion input and Conversion Complete output do not influence the actual analog-to-digital conversion in any way.

Internally the ADD3501 is always continuously converting the analog voltage present at its inputs. The Start Conversion input is used to control the transfer of information from the internal counter to the display latch.

An RS latch on the Start Conversion input allows a broad range of input pulse widths to be used on this signal. As shown in *Figure 3*, the Conversion Complete output goes to a logic "0" on the rising edge of the Start Conversion pulse and goes to a logic "1" some time later when the new conversion is transferred from the internal counter to the display latch. Since the Start Conversion pulse can occur at any time during the conversion cycle, the amount of time from Start Conversion to Conversion Complete will vary. The maximum time is $64{,}512 \times 1/f_{IN}$ and the minimum time is $256 \times 1/f_{IN}$.

Timing Waveforms



TL/H/5681–4

Figure 2. Conversion Cycle Timing Diagram for Free Running Operation

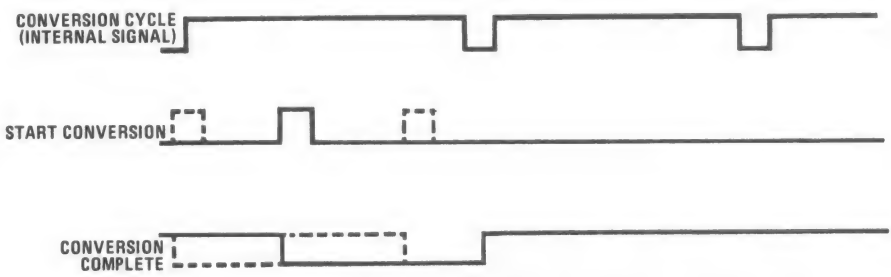


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Figure 3. Conversion Cycle Timing Diagram Operating with Start Conversion Input

Applications

SYSTEM DESIGN CONSIDERATIONS

Perhaps the most important thing to consider when designing a system using the ADD3501 is power supply noise on the V_{CC} and ground lines. Because a single power supply is used and currents in the 300 mA range are being switched, good circuit layout techniques cannot be overemphasized. Great care has been exercised in the design of the ADD3501 to minimize these problems but poor printed circuit layout can negate these features.

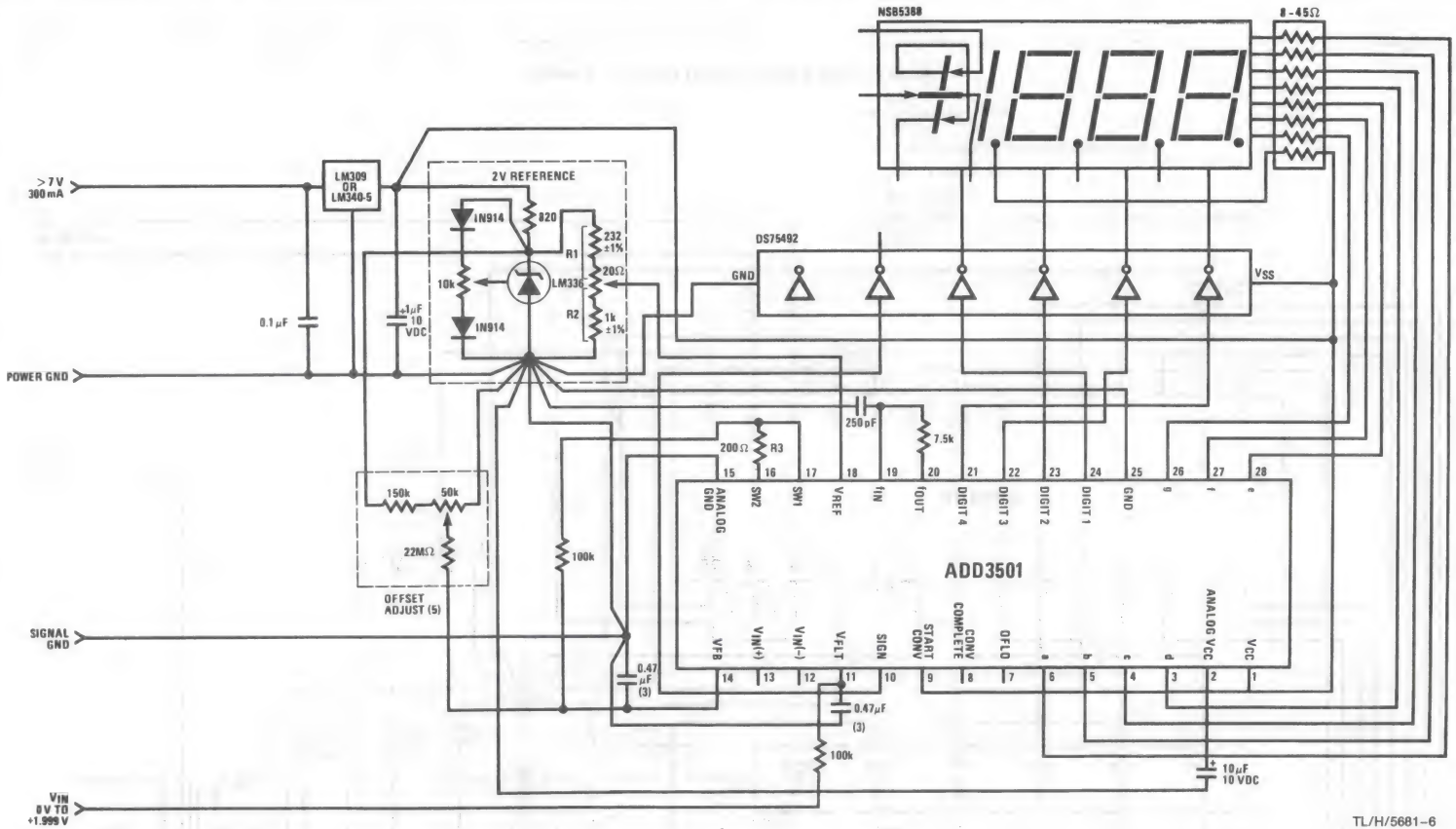
Figures 4, 5, and *6* show schematics of DVM systems. An attempt has been made to show, on these schematics, the proper distribution for ground and V_{CC}. To help isolate digital and analog portions of the circuit, the analog V_{CC} and ground have been separated from the digital V_{CC} and ground. Care must be taken to eliminate high current from flowing in the analog V_{CC} and ground wires. The most effective method of accomplishing this is to use a single ground point and a single V_{CC} point where all wires are brought together. In addition to this the conductors must be of sufficient size to prevent significant voltage drops.

To prevent switching noise from causing jitter problems, a voltage regulator with good high frequency response is necessary. The LM309 and the LM340-5 voltage regulators both function well and are shown in *Figures 4, 5,* and *6.* Adding more filtering than is shown will in general increase the jitter rather than decrease it. The most important characteristic of transients on the V_{CC} line is the duration of the transient and not its amplitude.

Figure 4 shows a DPM system which converts 0V to 1.999V operating from a non-isolated power supply. In this configuration the sign output could be + (logic "1") or − (logic "0") and it should be ignored. Higher voltages could be converted by placing a fixed divider on the input; lower voltages could be converted by placing a fixed divider on the feedback, as shown in *Figure 6.*

Figures 5 and *6* show systems operating with an isolated supply that will convert positive and negative inputs. 60 Hz common mode input becomes a problem in this configuration and a transformer with an electrostatic shield between primary and secondary windings is shown. The necessity for using a shielded transformer depends on the performance requirements and the actual application.

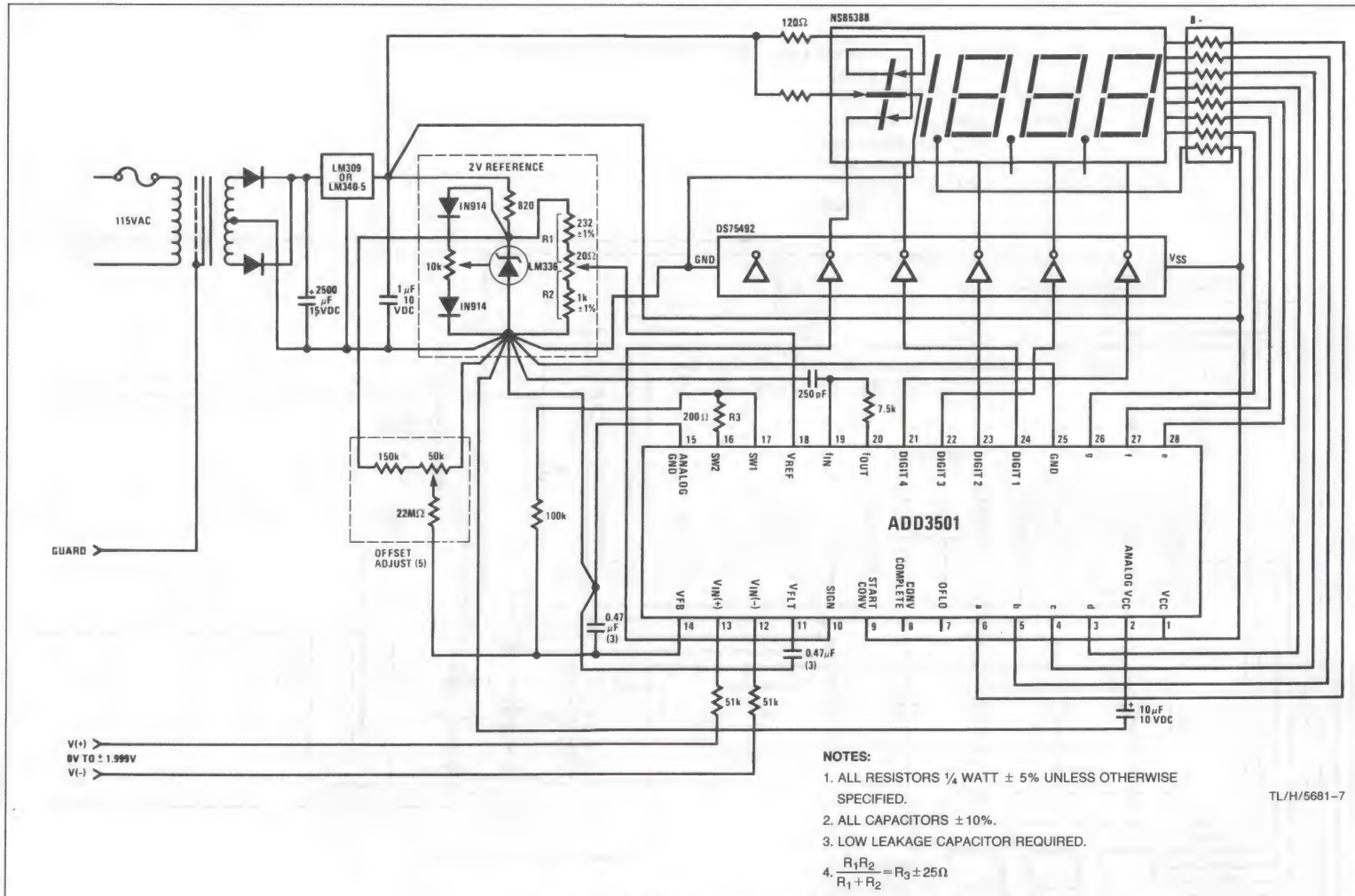
The filter capacitors connected to V_{FB} (pin 14) and V_{FLT} (pin 11) should be low leakage. In the application examples shown every 1.0nA of leakage current will cause 0.1mV error ($1.0\times10^{-9}A\times100k\Omega=0.1mV$). If the leakage current in both capacitors is exactly the same no error will result since the source impedances driving them are matched.



NOTES:

1. ALL RESISTORS ¼ WATT ± 5% UNLESS OTHERWISE SPECIFIED.
2. ALL CAPACITORS ±10%.
3. LOW LEAKAGE CAPACITOR REQUIRED.
4. $\frac{R_1R_2}{R_1+R_2}=R_3\pm25\Omega$

Figure 4. 3½-Digit DPM, + 1.999 Volts Full Scale



NOTES:

1. ALL RESISTORS ¼ WATT ± 5% UNLESS OTHERWISE SPECIFIED.
2. ALL CAPACITORS ±10%.
3. LOW LEAKAGE CAPACITOR REQUIRED.
4. $\frac{R_1R_2}{R_1+R_2}=R_3\pm25\Omega$

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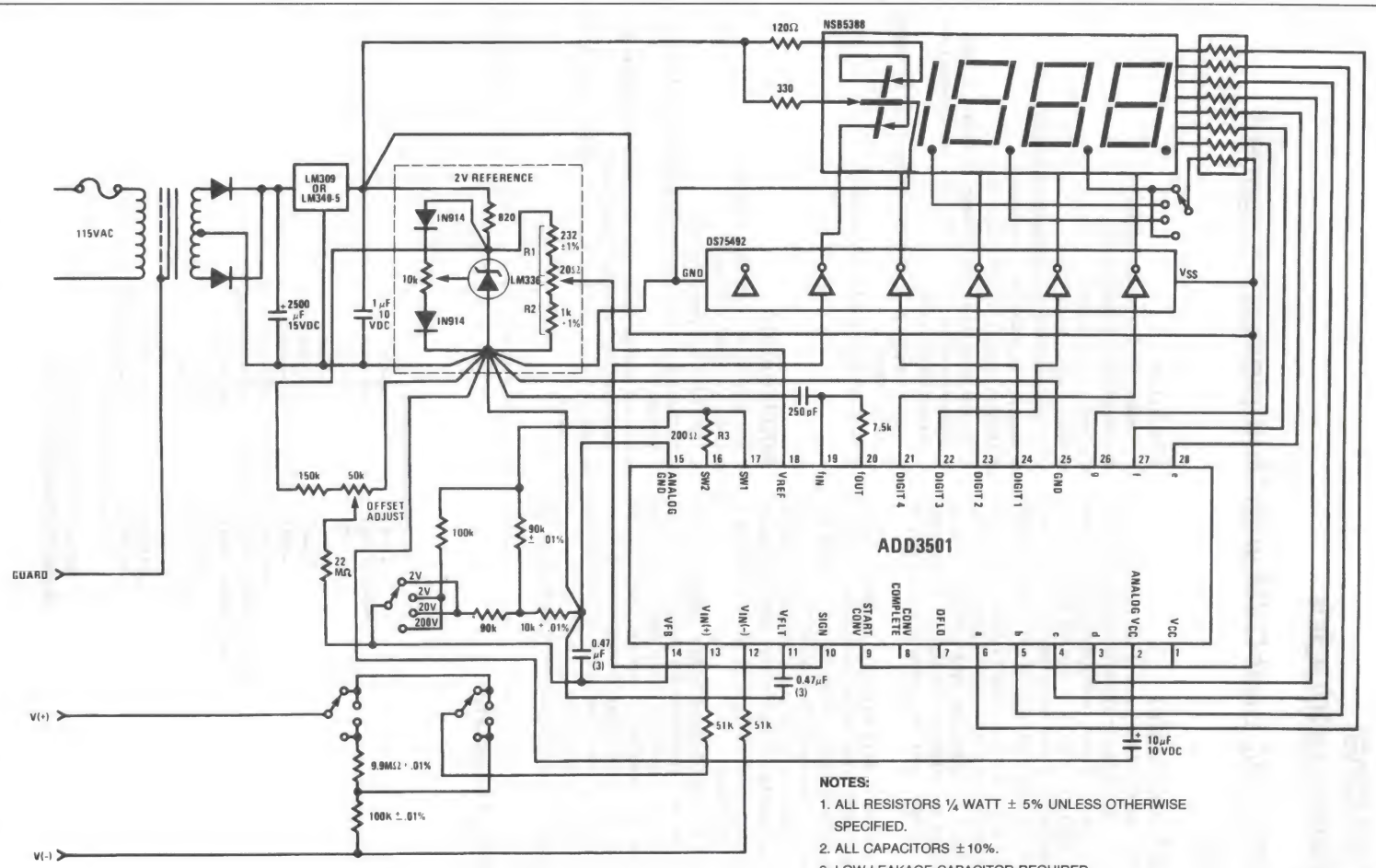
Figure 5. 3½-Digit DPM, ± 1.999 Volts Full Scale



NOTES:

1. ALL RESISTORS ¼ WATT ± 5% UNLESS OTHERWISE SPECIFIED.
2. ALL CAPACITORS ±10%.
3. LOW LEAKAGE CAPACITOR REQUIRED.
4. $\frac{R_1R_2}{R_1+R_2}=R_3\pm25\Omega$

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Figure 6. 3½-Digit DVM, Four Decade, ±0.2V, ±2V, ±20V and ±200V Full Scale

National Semiconductor

ADD3701 3¾ Digit DVM with Multiplexed 7-Segment Output

General Description

The ADD3701 (MM74C936-1) monolithic DVM circuit is manufactured using standard complementary MOS (CMOS) technology. A pulse modulation analog-to-digital conversion technique is used and requires no external precision components. In addition, this technique allows the use of a reference voltage that is the same polarity as the input voltage.

One 5V (TTL) power supply is required. Operating with an isolated supply allows the conversion of positive as well as negative voltages. The sign of the input voltage is automatically determined and output on the sign pin. If the power supply is not isolated, only one polarity of voltage may be converted.

The conversion rate is set by an internal oscillator. The frequency of the oscillator can be set by an external RC network or the oscillator can be driven from an external frequency source. When using the external RC network, a square wave output is available. It is important to note that great care has been taken to synchronize digit multiplexing with the A/D conversion timing to eliminate noise due to power supply transients.

The ADD3701 has been designed to drive 7-segment multiplexed LED displays directly with the aid of external digit buffers and segment resistors. Under condition of overrange, the overflow output will go high and the display will read +OFL or −OFL, depending on whether the input voltage is positive or negative. In addition to this, the most significant digit is blanked when zero.

A start conversion input and a conversion complete output are included.

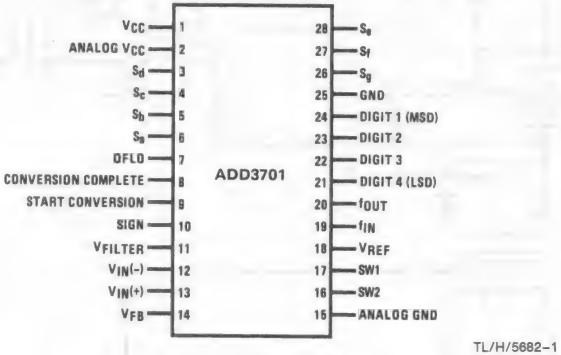
Features

- Operates from single 5V supply
- Converts 0 to ±3999 counts
- Multiplexed 7-segment
- Drives segments directly
- No external precision components necessary
- Accuracy specified over temperature
- Medium speed — 400 ms/conversion
- Internal clock set with RC network or driven externally
- Overrange indicated by +OFL or −OFL display reading and OFLO output
- Analog inputs in applications shown can withstand ±200 Volts
- ADD3701 equivalent to MM74C936-1

Applications

- Low cost digital power supply readouts
- Low cost digital multimeters
- Low cost digital panel meters
- Eliminate analog multiplexing by using remote A/D converters
- Convert analog transducers (temperature, pressure, displacement, etc.) to digital transducers
- Indicators and displays requiring readout up to 3999 counts

Connection Diagram



TL/H/5682–1

Order Number ADD3701CCN
See NS Package Number N28B

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Voltage at Any Pin except Start Conversion	$-0.3V$ to $V_{CC}+0.3V$
Voltage at Start Conversion	$-0.3V$ to $+15.0V$
ESD Susceptibility (Note 5)	TBDV
Operating Temperature Range (T_A)	$-40°C$ to $+85°C$
Package Dissipation at $T_A=25°C$	800mW
Operating V_{CC} Range	4.5V to 6.0V
Absolute Maximum V_{CC}	6.5V
Lead Temp. (Soldering, 10 seconds)	260°C
Storage Temperature Range	$-65°C$ to $+150°C$

Electrical Characteristics

$4.75V \le V_{CC} \le 5.25V$, $-40°C \le T_A \le +85°C$, unless otherwise specified.

Parameter		Conditions	Min	Typ[2]	Max	Units
$V_{IN(1)}$	Logical "1" Input Voltage		$V_{CC}-1.5$			V
$V_{IN(0)}$	Logical "0" Input Voltage				1.5	V
$V_{OUT(0)}$	Logical "0" Output Voltage (All Digital Outputs Except Digital Outputs)	$I_O=1.1$ mA			0.4	V
$V_{OUT(0)}$	Logical "0" Output Voltage (Digit Outputs)	$I_O=0.7$ mA			0.4	V
$V_{OUT(1)}$	Logical "1" Output Voltage (All Segment Outputs)	$I_O=50$ mA @ $T_J=25°C$ $V_{CC}=5V$ $I_O=30$ mA @ $T_J=100°C$	$V_{CC}-1.6$ $V_{CC}-1.6$	$V_{CC}-1.3$ $V_{CC}-1.3$		V V
$V_{OUT(1)}$	Logical "1" Output Voltage (All Digital Outputs Except Segment Outputs)	$I_O=500$ μA (Digit Outputs) $I_O=360$ μA (Conv. Complete, +/−, OFLO Outputs)	$V_{CC}-0.4$			V
I_{SOURCE}	Output Source Current (Digital Outputs)	$V_{OUT}=1.0$ V	2.0			mA
$I_{IN(1)}$	Logical "1" Input Current (Start Conversion)	$V_{IN}=15V$			1.0	μA
$I_{IN(0)}$	Logical "0" Input Current (Start Conversion)	$V_{IN}=0V$	−1.0			μA
I_{CC}	Supply Current	Segments and Digits Open		0.5	10	mA
f_{OSC}	Oscillator Frequency			0.6/RC		kHz
f_{IN}	Clock Frequency		100		640	kHz
f_C	Conversion Rate			$f_{IN}/129,024$		conv./sec
f_{MUX}	Digit Mux Rate			$f_{IN}/512$		Hz
t_{BLANK}	Inter Digit Blanking Time			$1/(32f_{MUX})$		seconds
t_{SCPW}	Start Conversion Pulse Width		200		DC	ns

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All typicals given for $T_A=25°C$.

Note 3: Full scale = 4000 counts; therefore 0.025% of full scale = 1 count and 0.05% of full scale = 2 counts.

Note 4: For 2.000 Volts full scale, 1 mV = 2 counts.

Note 5: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

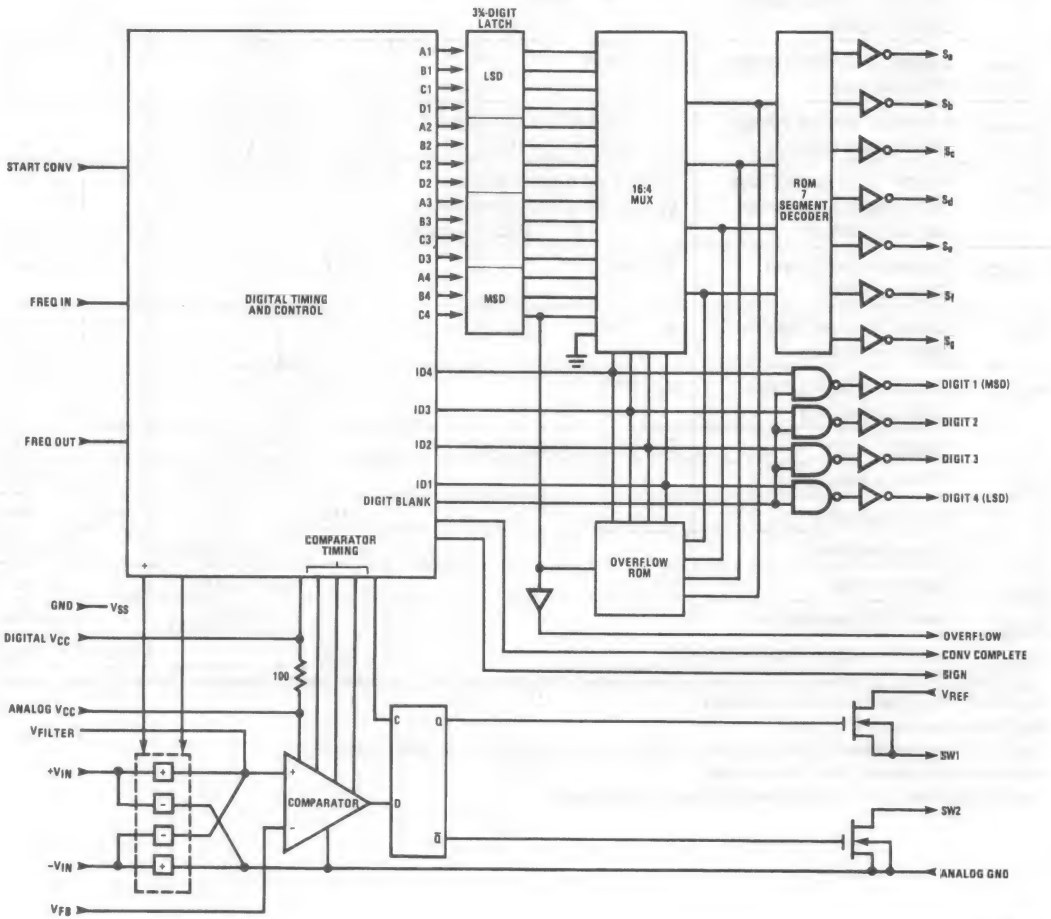
Electrical Characteristics (Continued)

$t_C = 2.5$ conversions/second, $0°C \leq T_A \leq +70°C$, unless otherwise specified.

Parameter	Conditions	Min	Typ[2]	Max	Units
Non-Linearity of Output Reading	$V_{IN} = 0 - 2V$ Full Scale $V_{IN} = 0 - 200$ mV Full Scale	−0.05 −0.05	±0.025 ±0.025	±0.05 ±0.05	% full scale (Note 3)
Quantization Error		−1		+0	counts
Offset Error, $V_{IN} = 0V$		−0.5	+1.5	+3	mV (Note 4)
Rollover Error		−0		+0	counts
Analog Input Current ($V_{IN}+$, $V_{IN}-$)	$T_A = 25°C$	−5	±1	+5	nA

Block Diagram

ADD3701 3¾-Digit DVM



TL/H/5682–2

Theory of Operation

A schematic for the analog loop is shown in *Figure 1.* The output of SW1 is either at V_{REF} or zero volts, depending on the state of the D flip-flop. If Q is at a high level, $V_{OUT} = V_{REF}$ and if Q is at a low level $V_{OUT} = 0V$. This voltage is then applied to the low pass filter comprised of R1 and C1. The output of this filter, V_{FB}, is connected to the negative input of the comparator, where it is compared to the analog input voltage, V_{IN}. The output of the comparator is connected to the D input of the D flip-flop. Information is then transferred from the D input to the Q and Q outputs on the positive edge of clock. This loop forms an oscillator whose duty cycle is precisely related to the analog input voltage, V_{IN}.

An example will demonstrate this relationship. Assume the input voltage is equal to 0.500 V. If the Q output of the D flip-flop is high then V_{OUT} will equal V_{REF} (2.000 V) and V_{FB} will charge toward 2 V with a time constant equal to R_1C_1. At some time V_{FB} will exceed 0.500 V and the comparator output will switch to 0V. At the next clock rising edge the Q output of the D flip-flop will switch to ground, causing V_{OUT} to switch to 0V. At this time V_{FB} will start discharging toward 0V with a time constant R_1C_1. When V_{FB} is less than 0.5 V the comparator output will switch high. On the rising edge of the next clock the Q output of the D flip-flop will switch high and the process will repeat. There exists at the output of SW1 a square wave pulse train with positive amplitude V_{REF} and negative amplitude 0V.

The DC value of this pulse train is:

$$V_{OUT} = V_{REF} \frac{t_{ON}}{t_{ON} + t_{OFF}} = V_{REF} \text{ (duty cycle)}$$

The lowpass filter will pass the DC value and then:

$$V_{FB} = V_{REF} \text{ (duty cycle)}$$

Since the closed loop system will always force V_{FB} to equal V_{IN}, we can then say that:

$$V_{IN} = V_{FB} = V_{REF} \text{ (duty cycle)}$$

or

$$\frac{V_{IN}}{V_{REF}} = \text{(duty cycle)}$$

The duty cycle is logically ANDed with the input frequency f_{IN}. The resultant frequency f equals:

$$f = \text{(duty cycle)} \times \text{(clock)}$$

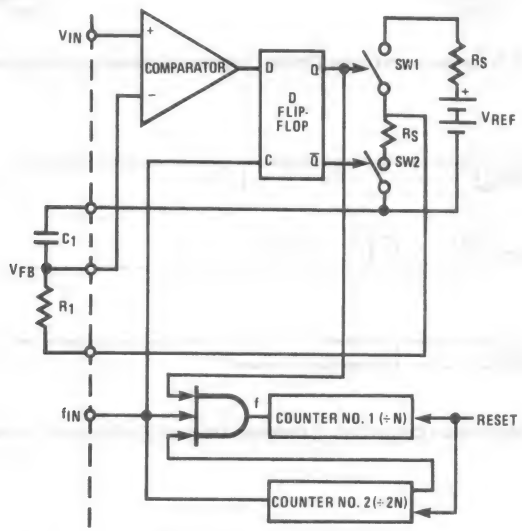
Frequency f is accumulated by counter no. 1 for a time determined by counter no. 2. The count contained in counter no. 1 is then:

$$\text{(count)} = \frac{f}{\text{(clock)}/N} = \frac{\text{(duty cycle)} \times \text{(clock)}}{\text{(clock)}/N}$$

$$= \frac{V_{IN}}{V_{REF}} \times N$$

For the ADD3701 $N = 4000$.

Schematic Diagram



$V_{IN} = V_{FB} = V_{REF} \times \text{(duty cycle)}$

$f = \text{(duty cycle)} \times f_{IN}$

$$\text{Count in Counter No. 1} = \frac{f}{f_{IN}/N} = \frac{\text{(duty cycle)} \times f_{IN}}{f_{IN}/N} = \frac{V_{IN}}{V_{REF}} \times N$$

FIGURE 1. Analog Loop Schematic Pulse Modulation A/D Converter

General Information

The timing diagram, shown in *Figure 2*, gives operation for the free running mode. Free running operation is obtained by connecting the Start Conversion input to logic "1" (V_{CC}). In this mode the analog input is continuously converted and the display is updated at a rate equal to $129{,}024 \times 1/f_{IN}$.

The rising edge of the Conversion Complete output indicates that new information has been transferred from the internal counter to the display latch. This information will remain in the display latch until the next low-to-high transition of the Conversion Complete output. A logic "1" will be maintained on the Conversion Complete output for a time equal to $128 \times 1/f_{IN}$.

Figure 3 gives the operation using the Start Conversion input. It is important to note that the Start Conversion input and Conversion Complete output do not influence the actual analog-to-digital conversion in any way.

Internally the ADD3701 is always continuously converting the analog voltage present at its input. The Start Conversion input is used to control the transfer of information from the internal counter to the display latch.

An RS latch on the Start Conversion input allows a broad range of input pulse widths to be used on this signal. As shown in *Figure 3*, the Conversion Complete output goes to a logic "0" on the rising edge of the Start Conversion pulse and goes to a logic "1" some time later when the new conversion is transferred from the internal counter to the display latch. Since the Start Conversion pulse can occur at any time during the conversion cycle, the amount of time from Start Conversion to Conversion Complete will vary. The maximum time is $129{,}024 \times 1/f_{IN}$ and the minimum time is $512 \times 1/f_{IN}$.

Timing Waveforms

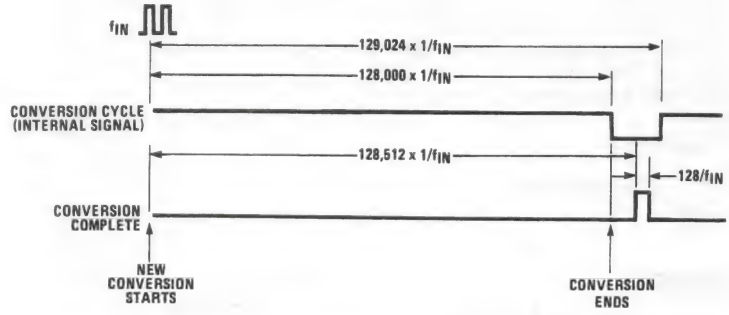


FIGURE 2. Conversion Cycle Timing Diagram for Free Running Operation

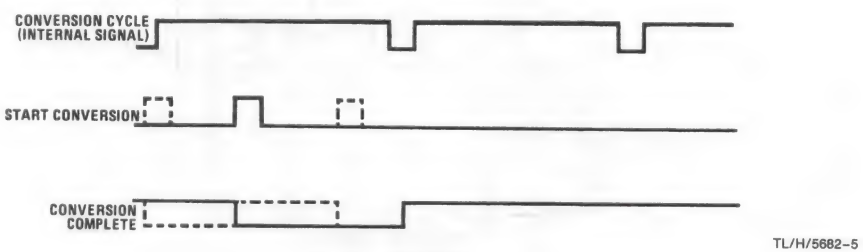


FIGURE 3. Conversion Cycle Timing Diagram Operating with Start Conversion Input

Applications

SYSTEM DESIGN CONSIDERATIONS

Perhaps the most important thing to consider when designing a system using the ADD3701 is power supply noise on the V_{CC} and ground lines. Because a single power supply is used and currents in the 300 mA range are being switched, good circuit layout techniques cannot be overemphasized. Great care has been exercised in the design of the ADD3701 to minimize these problems but poor printed circuit layout can negate these features.

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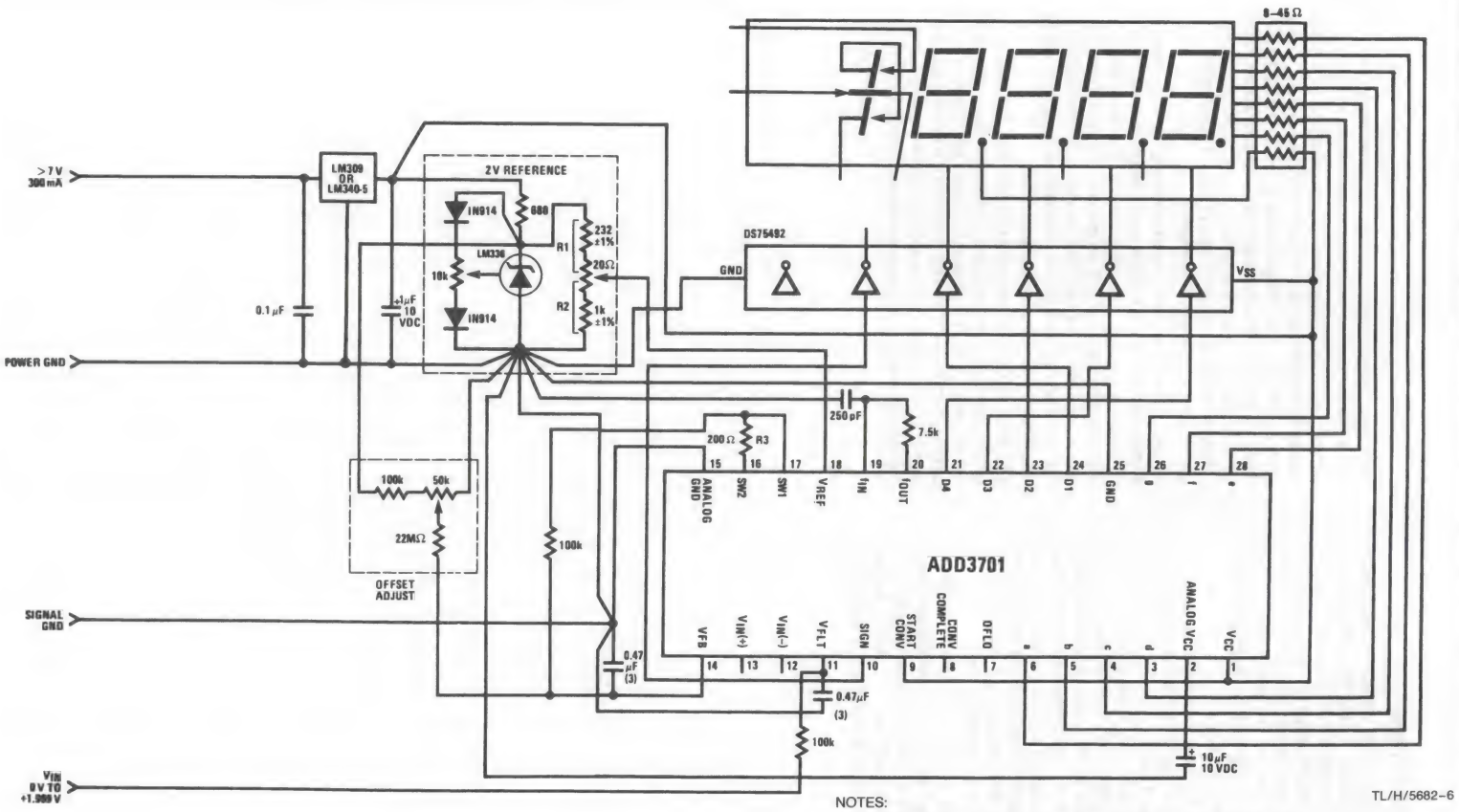
Figure 4 shows a DPM system which converts 0 to +3.999 counts operating from a non-isolated power supply. In this configuration the sign output could be + (logic "1") or − (logic "0") and it should be ignored. Higher voltages could be converted by placing a fixed divider on the input; lower voltages could be converted by placing a fixed divider on the feedback, as shown in *Figure 5*.

Figures 5 and *6* show systems operating with an isolated supply that will convert positive and negative inputs. 60 Hz common mode input becomes a problem in this configuration and a transformer with an electrostatic shield between primary and secondary windings is shown. The necessity for using a shielded transformer depends on the performance requirements and the actual application.

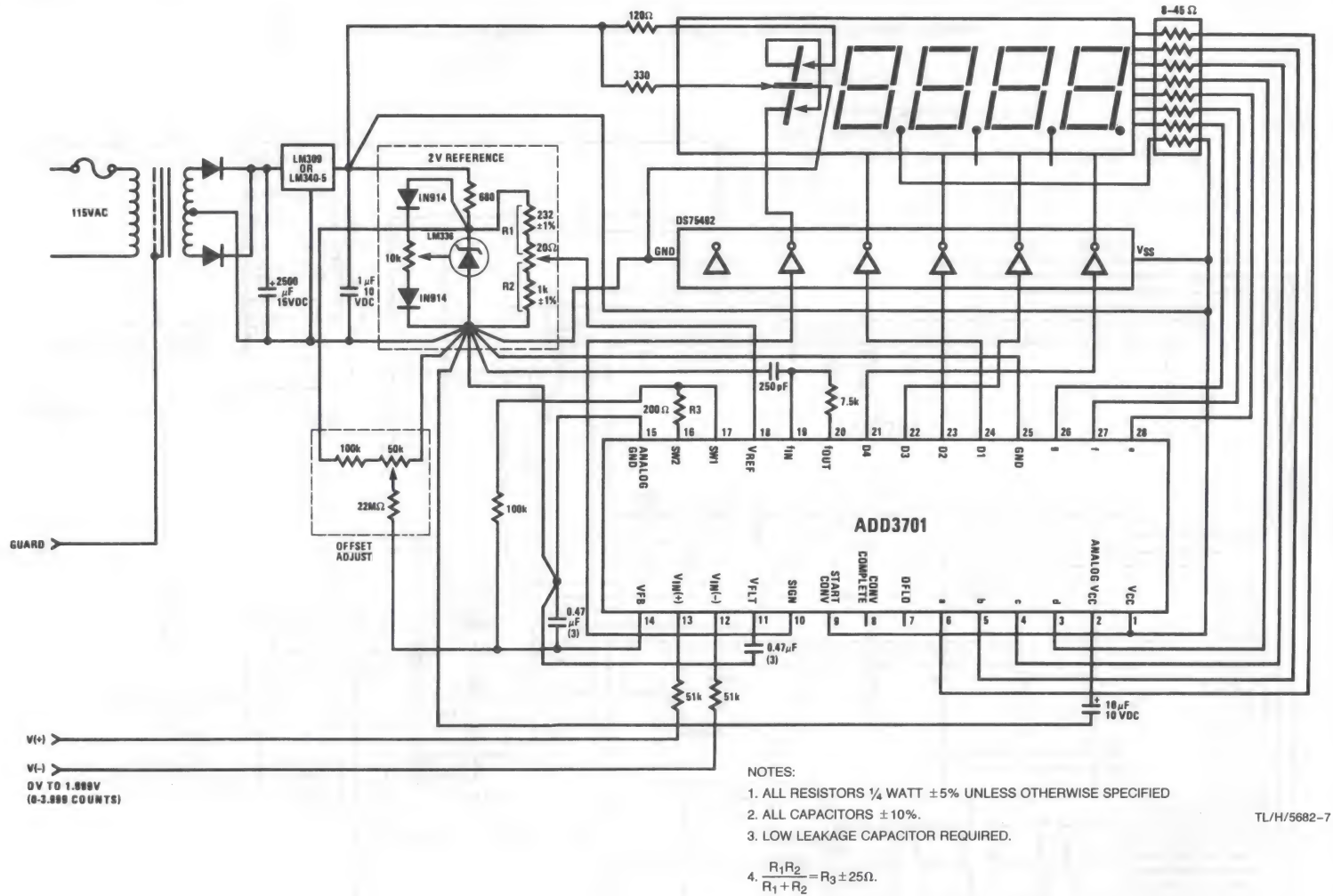
The filter capacitors connected to V_{FB} (pin 14) and V_{FLT} (pin 11) should be low leakage. In the application examples shown every 1.0 nA of leakage current will cause 0.1 mV error ($1.0 \times 10^{-9}A \times 100\ k\Omega = 0.1$ mV). If the leakage current in both capacitors is exactly the same no error will result since the source impedances driving them are matched.



NOTES:

1. ALL RESISTORS ¼ WATT ± 5% UNLESS OTHERWISE SPECIFIED
2. ALL CAPACITORS ±10%
3. LOW LEAKAGE CAPACITOR REQUIRED.
4. $\frac{R_1R_2}{R_1 + R_2} = R3 \pm 25\Omega$

Figure 4. 3¾-Digital DPM, +3.999 Count Full Scale

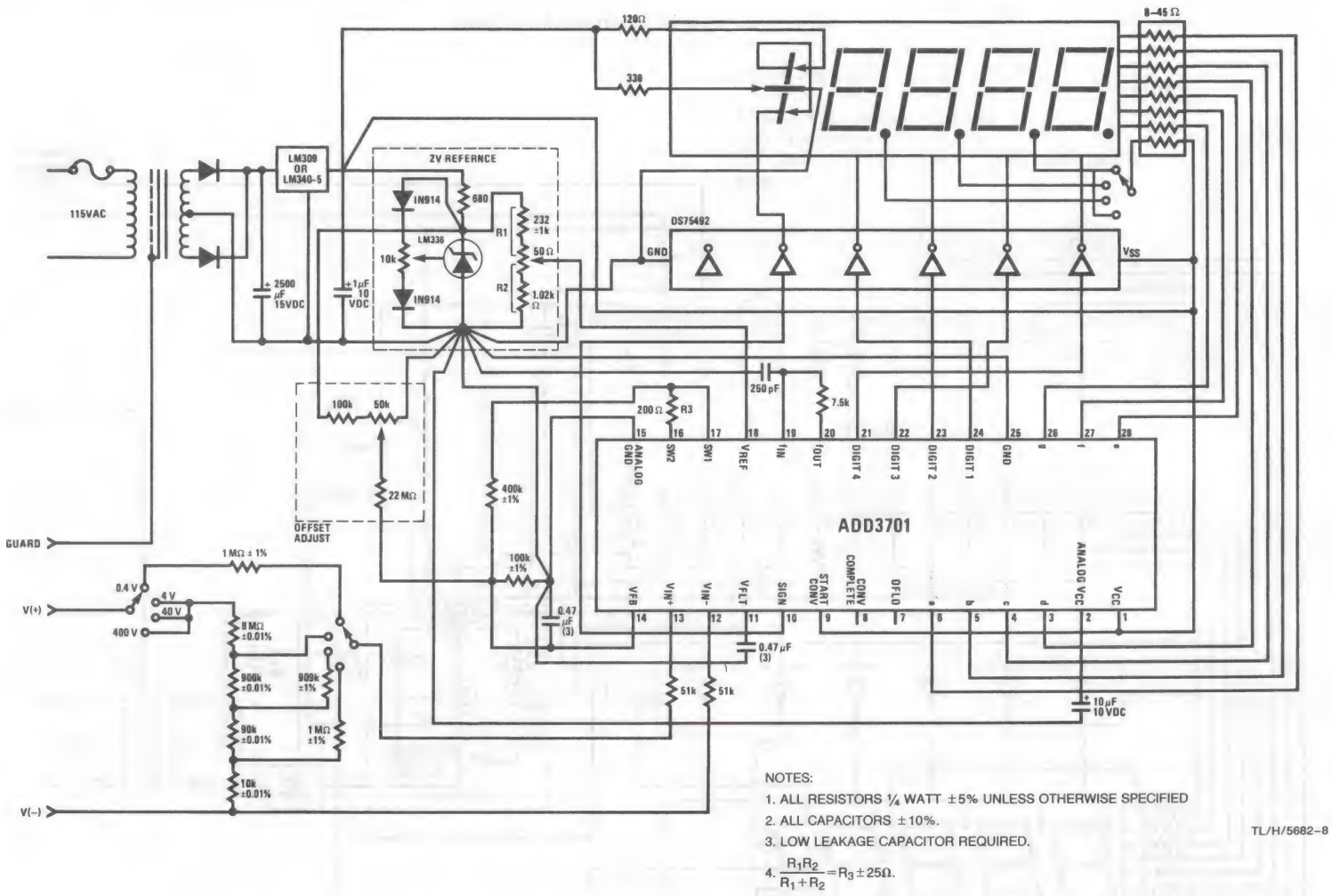


NOTES:

1. ALL RESISTORS ¼ WATT ±5% UNLESS OTHERWISE SPECIFIED
2. ALL CAPACITORS ±10%.
3. LOW LEAKAGE CAPACITOR REQUIRED.
4. $\frac{R_1R_2}{R_1+R_2}=R_3\pm25\Omega$.

TL/H/5682–7

Figure 5. 3¾-Digit DPM, ±3.999 Counts Full Scale

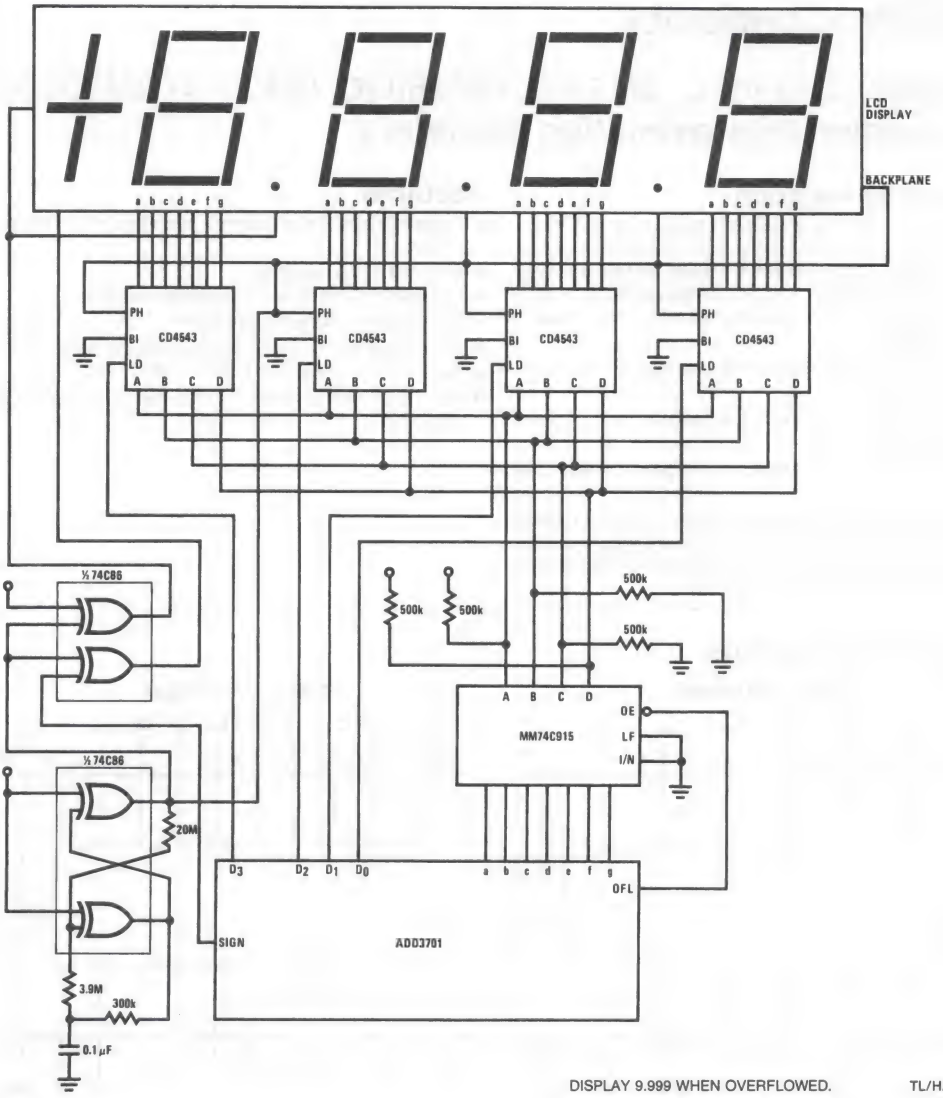


NOTES:

1. ALL RESISTORS ¼ WATT ±5% UNLESS OTHERWISE SPECIFIED
2. ALL CAPACITORS ±10%.
3. LOW LEAKAGE CAPACITOR REQUIRED.
4. $\frac{R_1R_2}{R_1+R_2}=R_3\pm25\Omega$.

TL/H/5682–8

Figure 6. 3¾-Digit DVM, Four Decade, ±0.4V, ±4V, ±40V, and ±400V Full Scale



DISPLAY 9.999 WHEN OVERFLOWED. ALL DIGITS CAN ALSO BE BLANKED AT OVERFLOW BY TYING OFL TO B1 ON THE CD4543s.

TL/H/5682–9

Figure 7. ADD3701 Driving Liquid Crystal Display

National Semiconductor

DM2502/DM2502C, DM2503/DM2503C, DM2504/DM2504C Successive Approximation Registers

General Description

The DM2502, DM2503 and DM2504 are 8-bit and 12-bit TTL registers designed for use in successive approximation A/D converters. These devices contain all the logic and control circuits necessary (in combination with a D/A converter) to perform successive approximation analog-to-digital conversions.

The DM2502 has 8 bits with serial capability and is not expandable.

The DM2503 has 8 bits and is expandable without serial capability.

The DM2504 has 12 bits with serial capability and expandability.

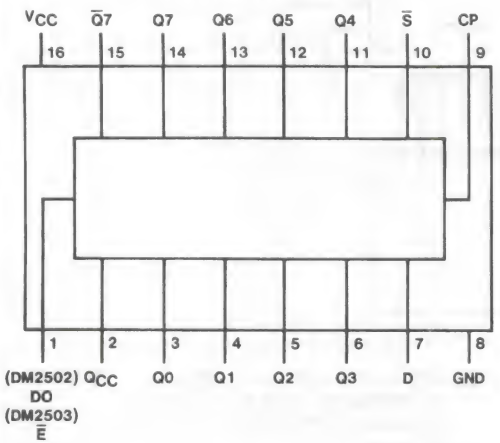
All three devices are available in ceramic DIP and molded Epoxy-B DIPs. The DM2502, DM2503 and DM2504 operate over −55°C to +125°C; the DM2502C, DM2503C and DM2504C operate over 0°C to +70°C.

Features

- Complete logic for successive approximation A/D converters
- 8-bit and 12-bit registers
- Capable of short cycle or expanded operation
- Continuous or start-stop operation
- Compatible with D/A converters using any logic code
- Active low or active high logic outputs
- Use as general purpose serial-to-parallel converter or ring counter

Connection Diagrams

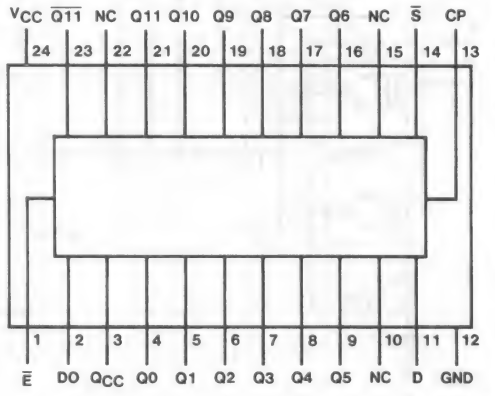
Dual-In-Line Package



TL/F/6612–1

Order Number DM2502J, DM2503J, DM2502CN or DM2503CN
See NS Package Number J16A or N16A

Dual-In-Line Package



TL/F/6612–2

Order Number DM2504J or DM2504CN
See NS Package Number J24A or N24A

See the LS/S/TTL Logic Databook for Complete Specifications

National Semiconductor

LM131A/LM131, LM231A/LM231, LM331A/LM331 Precision Voltage-to-Frequency Converters

General Description

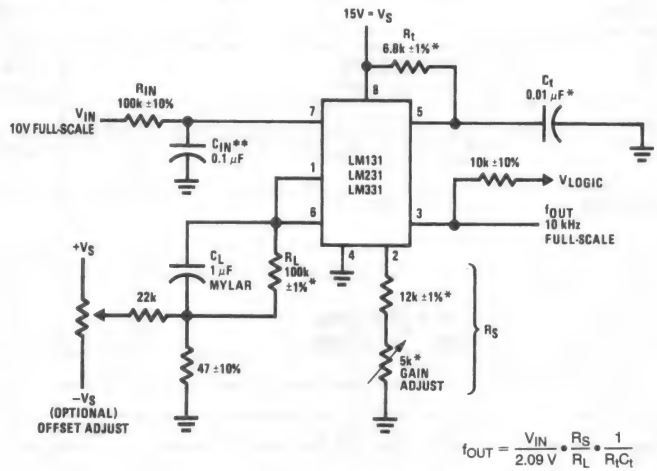
The LM131/LM231/LM331 family of voltage-to-frequency converters are ideally suited for use in simple low-cost circuits for analog-to-digital conversion, precision frequency-to-voltage conversion, long-term integration, linear frequency modulation or demodulation, and many other functions. The output when used as a voltage-to-frequency converter is a pulse train at a frequency precisely proportional to the applied input voltage. Thus, it provides all the inherent advantages of the voltage-to-frequency conversion techniques, and is easy to apply in all standard voltage-to-frequency converter applications. Further, the LM131A/LM231A/LM331A attains a new high level of accuracy versus temperature which could only be attained with expensive voltage-to-frequency modules. Additionally the LM131 is ideally suited for use in digital systems at low power supply voltages and can provide low-cost analog-to-digital conversion in microprocessor-controlled systems. And, the frequency from a battery powered voltage-to-frequency converter can be easily channeled through a simple photoisolator to provide isolation against high common mode levels.

The LM131/LM231/LM331 utilizes a new temperature-compensated band-gap reference circuit, to provide excellent accuracy over the full operating temperature range, at power supplies as low as 4.0V. The precision timer circuit has low bias currents without degrading the quick response necessary for 100 kHz voltage-to-frequency conversion. And the output is capable of driving 3 TTL loads, or a high voltage output up to 40V, yet is short-circuit-proof against V_{CC}.

Features

- Guaranteed linearity 0.01% max
- Improved performance in existing voltage-to-frequency conversion applications
- Split or single supply operation
- Operates on single 5V supply
- Pulse output compatible with all logic forms
- Excellent temperature stability, ±50 ppm/°C max
- Low power dissipation, 15 mW typical at 5V
- Wide dynamic range, 100 dB min at 10 kHz full scale frequency
- Wide range of full scale frequency, 1 Hz to 100 kHz
- Low cost

Typical Applications



*Use stable components with low temperature coefficients. See Typical Applications section.

**0.1μF or 1μF, See "Principles of Operation."

FIGURE 1. Simple Stand-Alone Voltage-to-Frequency Converter with ±0.03% Typical Linearity (f = 10 Hz to 11 kHz)

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/ Distributors for availability and specifications.

	LM131A/LM131	LM231A/LM231	LM331A/LM331
Supply Voltage	40V	40V	40V
Output Short Circuit to Ground	Continuous	Continuous	Continuous
Output Short Circuit to V_{CC}	Continuous	Continuous	Continuous
Input Voltage	−0.2V to +V_S	−0.2V to +V_S	−0.2V to +V_S
	T_{MIN} T_{MAX}	T_{MIN} T_{MAX}	T_{MIN} T_{MAX}
Operating Ambient Temperature Range	−55°C to +125°C	−25°C to +85°C	0°C to +70°C
Power Dissipation (P_D at 25°C) and Thermal Resistance (θ_{jA})			
(H Package) P_D	670 mW	570 mW	570 mW
θ_{jA}	150°C/W	150°C/W	150°C/W
(N Package) P_D		500 mW	500 mW
θ_{jA}		155°C/W	155°C/W
Lead Temperature (Soldering, 10 sec.)			
Dual-In-Line Package (Plastic)	260°C	260°C	260°C
Metal Can Package (TO-5)	260°C	260°C	260°C
ESD Susceptibility (Note 4)	TBD V	TBD V	TBD V

Electrical Characteristics $T_A = 25°C$ unless otherwise specified (Note 2)

Parameter	Conditions	Min	Typ	Max	Units
VFC Non-Linearity (Note 3)	4.5V ≤ V_S ≤ 20V		±0.003	±0.01	% Full-Scale
	$T_{MIN} \le T_A \le T_{MAX}$		±0.006	±0.02	% Full-Scale
VFC Non-Linearity In Circuit of *Figure 1*	V_S = 15V, f = 10 Hz to 11 kHz		±0.024	±0.14	%Full-Scale
Conversion Accuracy Scale Factor (Gain)	V_{IN} = −10V, R_S = 14 kΩ				
LM131, LM131A, LM231, LM231A		0.95	1.00	1.05	kHz/V
LM331, LM331A		0.90	1.00	1.10	kHz/V
Temperature Stability of Gain	$T_{MIN} \le T_A \le T_{MAX}$, 4.5V ≤ V_S ≤ 20V				
LM131/LM231/LM331			±30	±150	ppm/°C
LM131A/LM231A/LM331A			±20	±50	ppm/°C
Change of Gain with V_S	4.5V ≤ V_S ≤ 10V		0.01	0.1	%/V
	10V ≤ V_S ≤ 40V		0.006	0.06	%/V
Rated Full-Scale Frequency	V_{IN} = −10V	10.0			kHz
Gain Stability vs Time (1000 Hrs)	$T_{MIN} \le T_A \le T_{MAX}$		±0.02		% Full-Scale
Overrange (Beyond Full-Scale) Frequency	V_{IN} = −11V	10			%
INPUT COMPARATOR					
Offset Voltage			±3	±10	mV
LM131/LM231/LM331	$T_{MIN} \le T_A \le T_{MAX}$		±4	±14	mV
LM131A/LM231A/LM331A	$T_{MIN} \le T_A \le T_{MAX}$		±3	±10	mV
Bias Current			−80	−300	nA
Offset Current			±8	±100	nA
Common-Mode Range	$T_{MIN} \le T_A \le T_{MAX}$	−0.2		V_{CC}−2.0	V
TIMER					
Timer Threshold Voltage, Pin 5		0.63	0.667	0.70	× V_S
Input Bias Current, Pin 5	V_S = 15V				
All Devices	0V ≤ $V_{PIN\ 5}$ ≤ 9.9V		±10	±100	nA
LM131/LM231/LM331	$V_{PIN\ 5}$ = 10V		200	1000	nA
LM131A/LM231A/LM331A	$V_{PIN\ 5}$ = 10V		200	500	nA
$V_{SAT\ PIN\ 5}$ (Reset)	I = 5 mA		0.22	0.5	V

Electrical Characteristics $T_A = 25°C$ unless otherwise specified (Note 2) (Continued)

Parameter	Conditions	Min	Typ	Max	Units
CURRENT SOURCE (Pin 1)					
Output Current	$R_S = 14\ k\Omega$, $V_{PIN\ 1} = 0$				
LM131, LM131A, LM231, LM231A		126	135	144	μA
LM331, LM331A		116	136	156	μA
Change with Voltage	$0V \le V_{PIN\ 1} \le 10V$		0.2	1.0	μA
Current Source OFF Leakage					
LM131, LM131A			0.01	1.0	nA
LM231, LM231A, LM331, LM331A			0.02	10.0	nA
All Devices	$T_A = T_{MAX}$		2.0	50.0	nA
Operating Range of Current (Typical)			(10 to 500)		μA
REFERENCE VOLTAGE (Pin 2)					
LM131, LM131A, LM231, LM231A		1.76	1.89	2.02	V_{DC}
LM331, LM331A		1.70	1.89	2.08	V_{DC}
Stability vs Temperature			±60		ppm/°C
Stability vs Time, 1000 Hours			±0.1		%
LOGIC OUTPUT (Pin 3)					
V_{SAT}	$I = 5$ mA		0.15	0.50	V
	$I = 3.2$ mA (2 TTL Loads), $T_{MIN} \le T_A \le T_{MAX}$		0.10	0.40	V
OFF Leakage			±0.05	1.0	μA
SUPPLY CURRENT					
LM131, LM131A, LM231,	$V_S = 5V$	2.0	3.0	4.0	mA
LM231A	$V_S = 40V$	2.5	4.0	6.0	mA
LM331, LM331A	$V_S = 5V$	1.5	3.0	6.0	mA
	$V_S = 40V$	2.0	4.0	8.0	mA

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All specifications apply in the circuit of *Figure 3*, with $4.0V \le V_S \le 40V$, unless otherwise noted.

Note 3: Nonlinearity is defined as the deviation of f_{OUT} from $V_{IN} \times (10\ kHz/-10\ V_{DC})$ when the circuit has been trimmed for zero error at 10 Hz and at 10 kHz, over the frequency range 1 Hz to 11 kHz. For the timing capacitor, C_T, use NPO ceramic, Teflon®, or polystyrene.

Note 4: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Functional Block Diagram

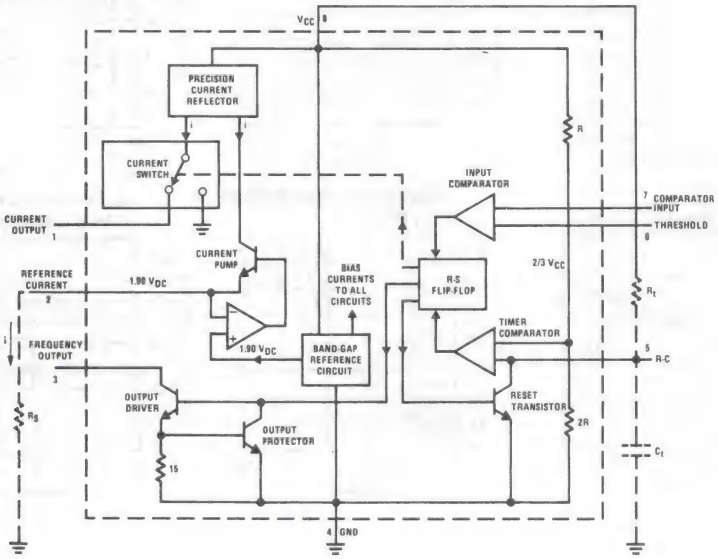


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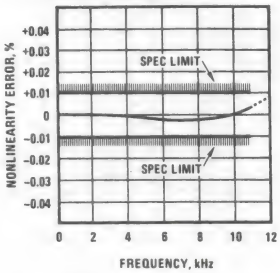
FIGURE 1a

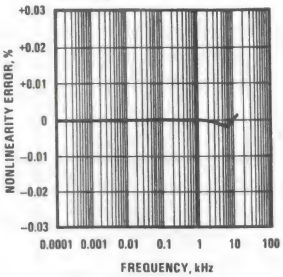
Typical Performance Characteristics

(All electrical characteristics apply for the circuit of *Figure 3*, unless otherwise noted.)

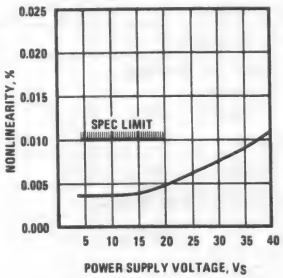
Nonlinearity Error, LM131 Family, as Precision V-to-F Converter (*Figure 3*)



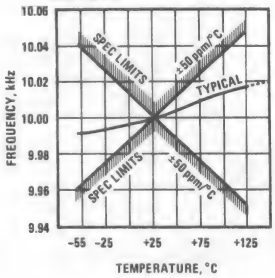
Nonlinearity Error, LM131 Family



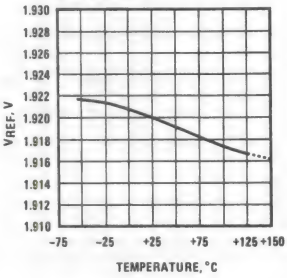
Nonlinearity vs Power Supply Voltage



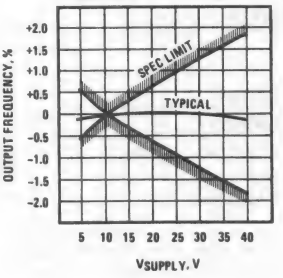
Frequency vs Temperature, LM131A



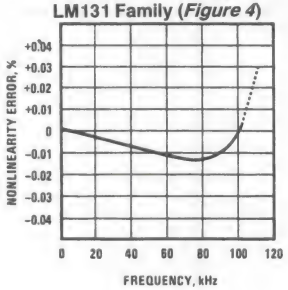
V_{REF} vs Temperature, LM131A



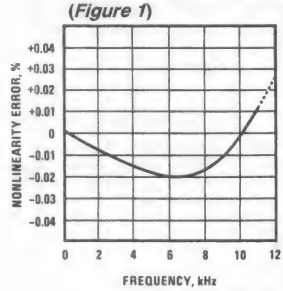
Output Frequency vs V_{SUPPLY}



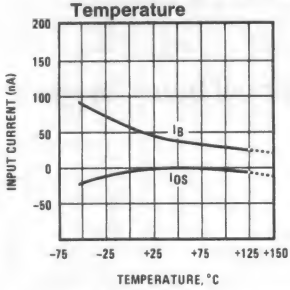
100 kHz Nonlinearity Error, LM131 Family (*Figure 4*)



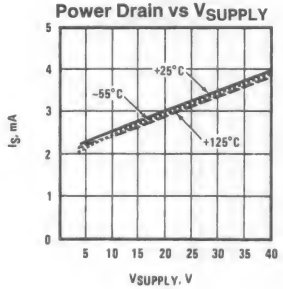
Nonlinearity Error, LM131 (*Figure 1*)



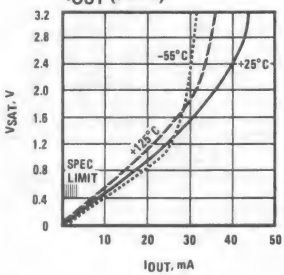
Input Current (Pins 6, 7) vs Temperature



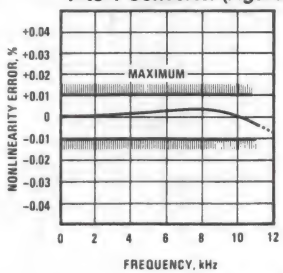
Power Drain vs V_{SUPPLY}



Output Saturation Voltage vs I_{OUT} (Pin 3)



Nonlinearity Error, Precision F-to-V Converter (*Figure 6*)



TL/H/5680–3

Typical Applications (Continued)

PRINCIPLES OF OPERATION OF A SIMPLIFIED VOLTAGE-TO-FREQUENCY CONVERTER

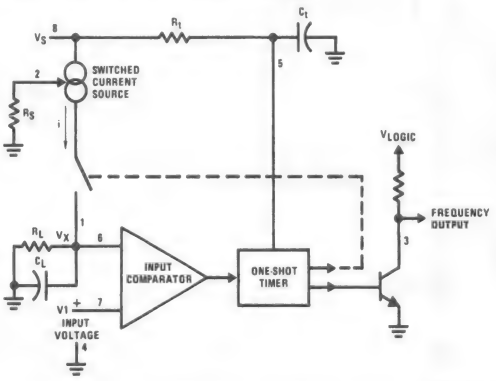
The LM131 is a monolithic circuit designed for accuracy and versatile operation when applied as a voltage-to-frequency (V-to-F) converter or as a frequency-to-voltage (F-to-V) converter. A simplified block diagram of the LM131 is shown in *Figure 2* and consists of a switched current source, input comparator, and 1-shot timer.

The operation of these blocks is best understood by going through the operating cycle of the basic V-to-F converter, *Figure 2*, which consists of the simplified block diagram of the LM131 and the various resistors and capacitors connected to it.

The voltage comparator compares a positive input voltage, V1, at pin 7 to the voltage, V_X, at pin 6. If V1 is greater, the comparator will trigger the 1-shot timer. The output of the timer will turn ON both the frequency output transistor and the switched current source for a period $t = 1.1\ R_tC_t$. During this period, the current i will flow out of the switched current source and provide a fixed amount of charge, $Q = i \times t$, into the capacitor, C_L. This will normally charge V_X up to a higher level than V1. At the end of the timing period, the current i will turn OFF, and the timer will reset itself.

Now there is no current flowing from pin 1, and the capacitor C_L will be gradually discharged by R_L until V_X falls to the level of V1. Then the comparator will trigger the timer and start another cycle.

The current flowing into C_L is exactly $I_{AVE} = i \times (1.1 \times R_tC_t) \times f$, and the current flowing out of C_L is exactly $V_X/R_L \cong V_{IN}/R_L$. If V_{IN} is doubled, the frequency will double to maintain this balance. Even a simple V-to-F converter can provide a frequency precisely proportional to its input voltage over a wide range of frequencies.



TL/H/5680–4

FIGURE 2. Simplified Block Diagram of Stand-Alone Voltage-to-Frequency Converter Showing LM131 and External Components

DETAIL OF OPERATION, FUNCTIONAL BLOCK DIAGRAM (*FIGURE 1a*)

The block diagram shows a band gap reference which provides a stable 1.9 V_{DC} output. This 1.9 V_{DC} is well regulated over a V_S range of 3.9V to 40V. It also has a flat, low temperature coefficient, and typically changes less than ½% over a 100°C temperature change.

The current pump circuit forces the voltage at pin 2 to be at 1.9V, and causes a current $i = 1.90V/R_S$ to flow. For $R_s = 14k$, $i = 135\ \mu A$. The precision current reflector provides a current equal to i to the current switch. The current switch switches the current to pin 1 or to ground depending on the state of the R_S flip-flop.

The timing function consists of an R_S flip-flop, and a timer comparator connected to the external R_tC_t network. When the input comparator detects a voltage at pin 7 higher than pin 6, it sets the R_S flip-flop which turns ON the current switch and the output driver transistor. When the voltage at pin 5 rises to ⅔ V_{CC}, the timer comparator causes the R_S flip-flop to reset. The reset transistor is then turned ON and the current switch is turned OFF.

However, if the input comparator still detects pin 7 higher than pin 6 when pin 5 crosses ⅔ V_{CC}, the flip-flop will not be reset, and the current at pin 1 will continue to flow, in its attempt to make the voltage at pin 6 higher than pin 7. This condition will usually apply under start-up conditions or in the case of an overload voltage at signal input. It should be noted that during this sort of overload, the output frequency will be 0; as soon as the signal is restored to the working range, the output frequency will be resumed.

The output driver transistor acts to saturate pin 3 with an ON resistance of about 50Ω. In case of overvoltage, the output current is actively limited to less than 50 mA.

The voltage at pin 2 is regulated at 1.90 V_{DC} for all values of i between 10 μA to 500 μA. It can be used as a voltage reference for other components, but care must be taken to ensure that current is not taken from it which could reduce the accuracy of the converter.

PRINCIPLES OF OPERATION OF BASIC VOLTAGE-TO-FREQUENCY CONVERTER (*FIGURE 1*)

The simple stand-alone V-to-F converter shown in *Figure 1* includes all the basic circuitry of *Figure 2* plus a few components for improved performance.

A resistor, $R_{IN} = 100\ k\Omega \pm 10\%$, has been added in the path to pin 7, so that the bias current at pin 7 (−80 nA typical) will cancel the effect of the bias current at pin 6 and help provide minimum frequency offset.

The resistance R_S at pin 2 is made up of a 12 kΩ fixed resistor plus a 5 kΩ (cermet, preferably) gain adjust rheostat. The function of this adjustment is to trim out the gain tolerance of the LM131, and the tolerance of R_t, R_L and C_t.

Typical Applications (Continued)

For best results, all the components should be stable low-temperature-coefficient components, such as metal-film resistors. The capacitor should have low dielectric absorption; depending on the temperature characteristics desired, NPO ceramic, polystyrene, Teflon or polypropylene are best suited.

A capacitor C_{IN} is added from pin 7 to ground to act as a filter for V_{IN}. A value of 0.01 μF to 0.1 μF will be adequate in most cases; however, in cases where better filtering is required, a 1 μF capacitor can be used. When the RC time constants are matched at pin 6 and pin 7, a voltage step at V_{IN} will cause a step change in f_{OUT}. If C_{IN} is much less than C_L, a step at V_{IN} may cause f_{OUT} to stop momentarily.

A 47Ω resistor, in series with the 1 μF C_L, is added to give hysteresis effect which helps the input comparator provide the excellent linearity (0.03% typical).

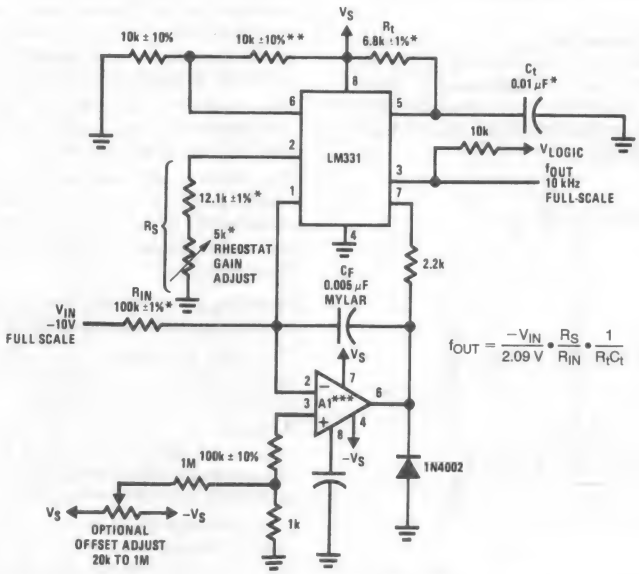
DETAIL OF OPERATION OF PRECISION V-TO-F CONVERTER (*FIGURE 3*)

In this circuit, integration is performed by using a conventional operational amplifier and feedback capacitor, C_F. When the integrator's output crosses the nominal threshold level at pin 6 of the LM131, the timing cycle is initiated.

The average current fed into the op amp's summing point (pin 2) is $i \times (1.1\ R_tC_t) \times f$ which is perfectly balanced with $-V_{IN}/R_{IN}$. In this circuit, the voltage offset of the LM131 input comparator does not affect the offset or accuracy of the V-to-F converter as it does in the stand-alone V-to-F converter; nor does the LM131 bias current or offset current. Instead, the offset voltage and offset current of the operational amplifier are the only limits on how small the signal can be accurately converted. Since op amps with voltage offset well below 1 mV and offset currents well below 2 nA are available at low cost, this circuit is recommended for best accuracy for small signals. This circuit also responds immediately to any change of input signal (which a stand-alone circuit does not) so that the output frequency will be an accurate representation of V_{IN}, as quickly as 2 output pulses' spacing can be measured.

In the precision mode, excellent linearity is obtained because the current source (pin 1) is always at ground potential and that voltage does not vary with V_{IN} or f_{OUT}. (In the stand-alone V-to-F converter, a major cause of non-linearity is the output impedance at pin 1 which causes i to change as a function of V_{IN}).

The circuit of *Figure 4* operates in the same way as *Figure 3*, but with the necessary changes for high speed operation.



TL/H/5680–5

*Use stable components with low temperature coefficients. See Typical Applications section.

**This resistor can be 5 kΩ or 10 kΩ for V_S = 8V to 22V, but must be 10 kΩ for V_S = 4.5V to 8V.

***Use low offset voltage and low offset current op amps for A1: recommended types LM108, LM308A, LF411A

FIGURE 3. Standard Test Circuit and Applications Circuit, Precision Voltage-to-Frequency Converter

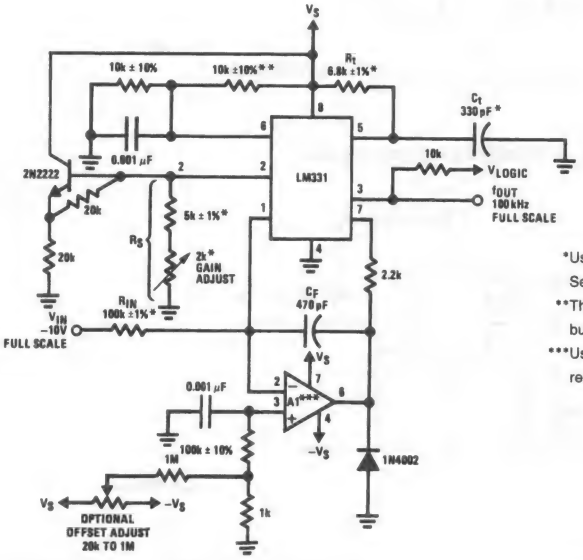
Typical Applications (Continued)

DETAILS OF OPERATION, FREQUENCY-TO-VOLTAGE CONVERTERS *(FIGURES 5 AND 6)*

In these applications, a pulse input at f_{IN} is differentiated by a C-R network and the negative-going edge at pin 6 causes the input comparator to trigger the timer circuit. Just as with a V-to-F converter, the average current flowing out of pin 1 is $I_{AVERAGE} = i \times (1.1\ R_tC_t) \times f$.

In the simple circuit of *FIGURE 5*, this current is filtered in the network $R_L = 100\ k\Omega$ and 1 μF. The ripple will be less than 10 mV peak, but the response will be slow, with a 0.1 second time constant, and settling of 0.7 second to 0.1% accuracy.

In the precision circuit, an operational amplifier provides a buffered output and also acts as a 2-pole filter. The ripple will be less than 5 mV peak for all frequencies above 1 kHz, and the response time will be much quicker than in *Figure 5*. However, for input frequencies below 200 Hz, this circuit will have worse ripple than *Figure 5*. The engineering of the filter time-constants to get adequate response and small enough ripple simply requires a study of the compromises to be made. Inherently, V-to-F converter response can be fast, but F-to-V response can not.



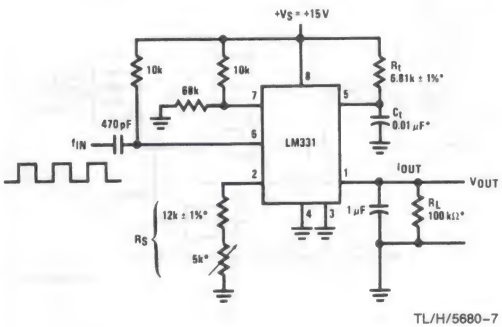
*Use stable components with low temperature coefficients. See Typical Applications section.

**This resistor can be 5 kΩ or 10 kΩ for V_S = 8V to 22V, but must be 10 kΩ for V_S = 4.5V to 8V.

***Use low offset voltage and low offset current op amps for A1: recommended types LF411A or LF356.

TL/H/5680–6

FIGURE 4. Precision Voltage-to-Frequency Converter, 100 kHz Full-Scale, ±0.03% Non-Linearity

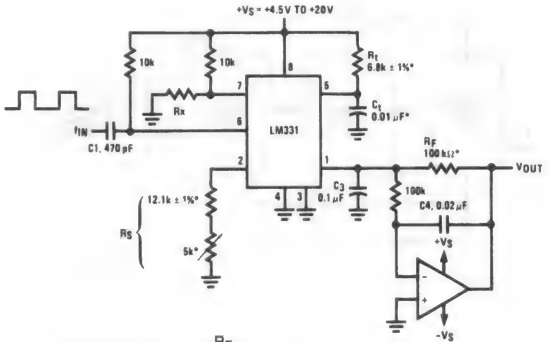


TL/H/5680–7

$$V_{OUT} = f_{IN} \times 2.09V \times \frac{R_L}{R_S} \times (R_tC_t)$$

*Use stable components with low temperature coefficients.

FIGURE 5. Simple Frequency-to-Voltage Converter, 10 kHz Full-Scale, ±0.06% Non-Linearity



TL/H/5680–8

$$V_{OUT} = -f_{IN} \times 2.09V \times \frac{R_F}{R_S} \times (R_tC_t)$$

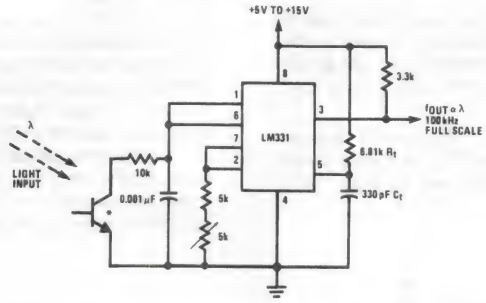
$$\text{SELECT Rx} = \frac{(V_S - 2V)}{0.2\ mA}$$

*Use stable components with low temperature coefficients.

FIGURE 6. Precision Frequency-to-Voltage Converter, 10 kHz Full-Scale with 2-Pole Filter, ±0.01% Non-Linearity Maximum

Typical Applications (Continued)

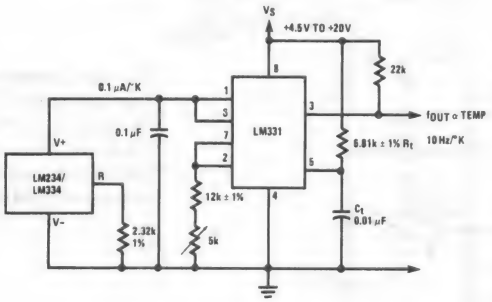
Light Intensity to Frequency Converter



TL/H/5680–9

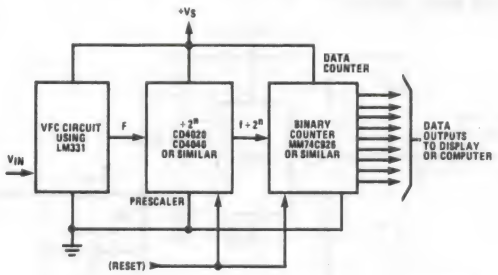
*L14F-1, L14G-1 or L14H-1, photo transistor (General Electric Co.) or similar

Temperature to Frequency Converter



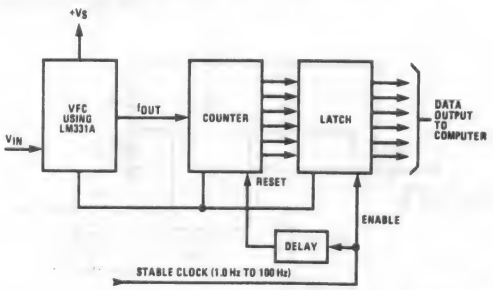
TL/H/5680–10

Long-Term Digital Integrator Using VFC



TL/H/5680–11

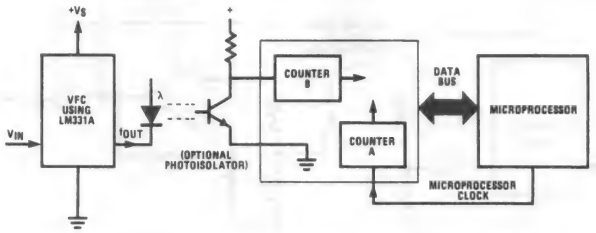
Basic Analog-to-Digital Converter Using Voltage-to-Frequency Converter



TL/H/5680–12

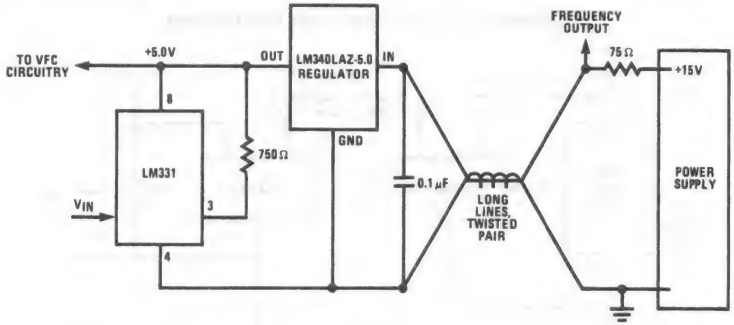
Typical Applications (Continued)

Analog-to-Digital Converter with Microprocessor



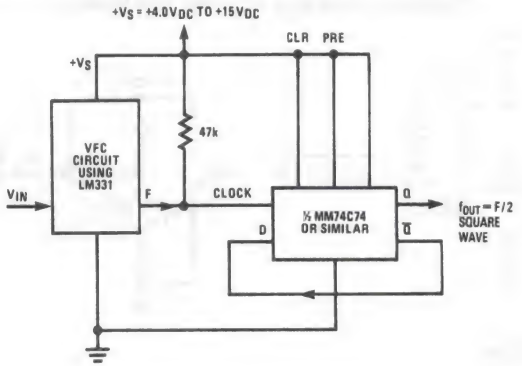
TL/H/5680–13

Remote Voltage-to-Frequency Converter with 2-Wire Transmitter and Receiver



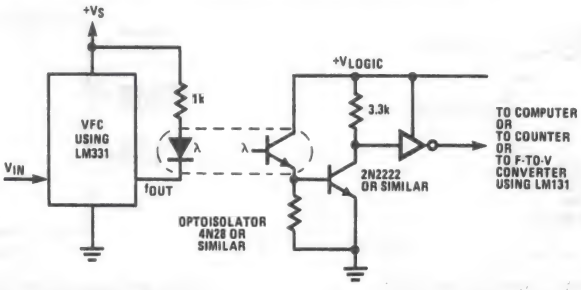
TL/H/5680–14

Voltage-to-Frequency Converter with Square-Wave Output Using ÷ 2 Flip-Flop



TL/H/5680–15

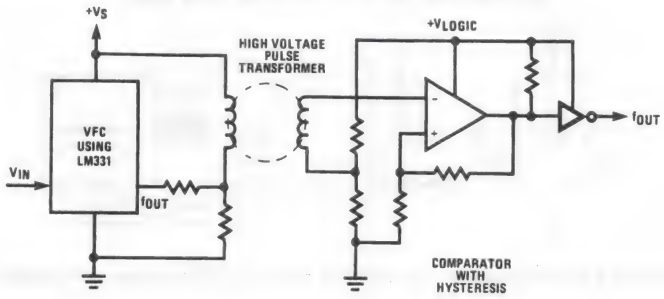
Voltage-to-Frequency Converter with Isolators



TL/H/5680–16

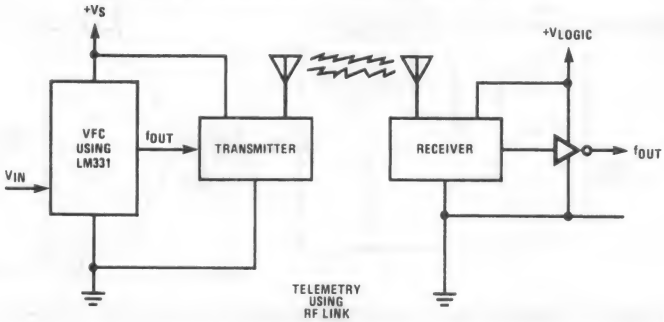
Typical Applications (Continued)

Voltage-to-Frequency Converter with Isolators



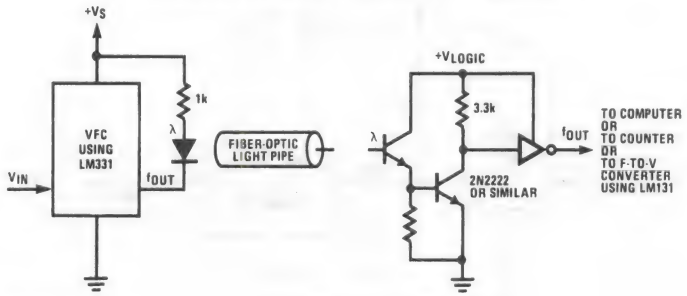
TL/H/5680–17

Voltage-to-Frequency Converter with Isolators



TL/H/5680–18

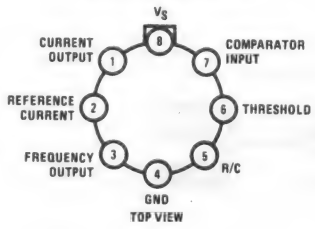
Voltage-to-Frequency Converter with Isolators



TL/H/5680–19

Connection Diagrams

Metal Can Package

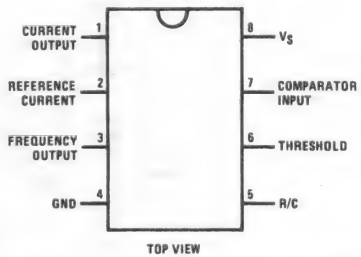


TL/H/5680–20

Note: Metal case is connected to pin 4 (GND.)

Order Number LM131AH, LM131H, LM231AH, LM231H, LM331AH or LM331H
See NS Package Number H08C

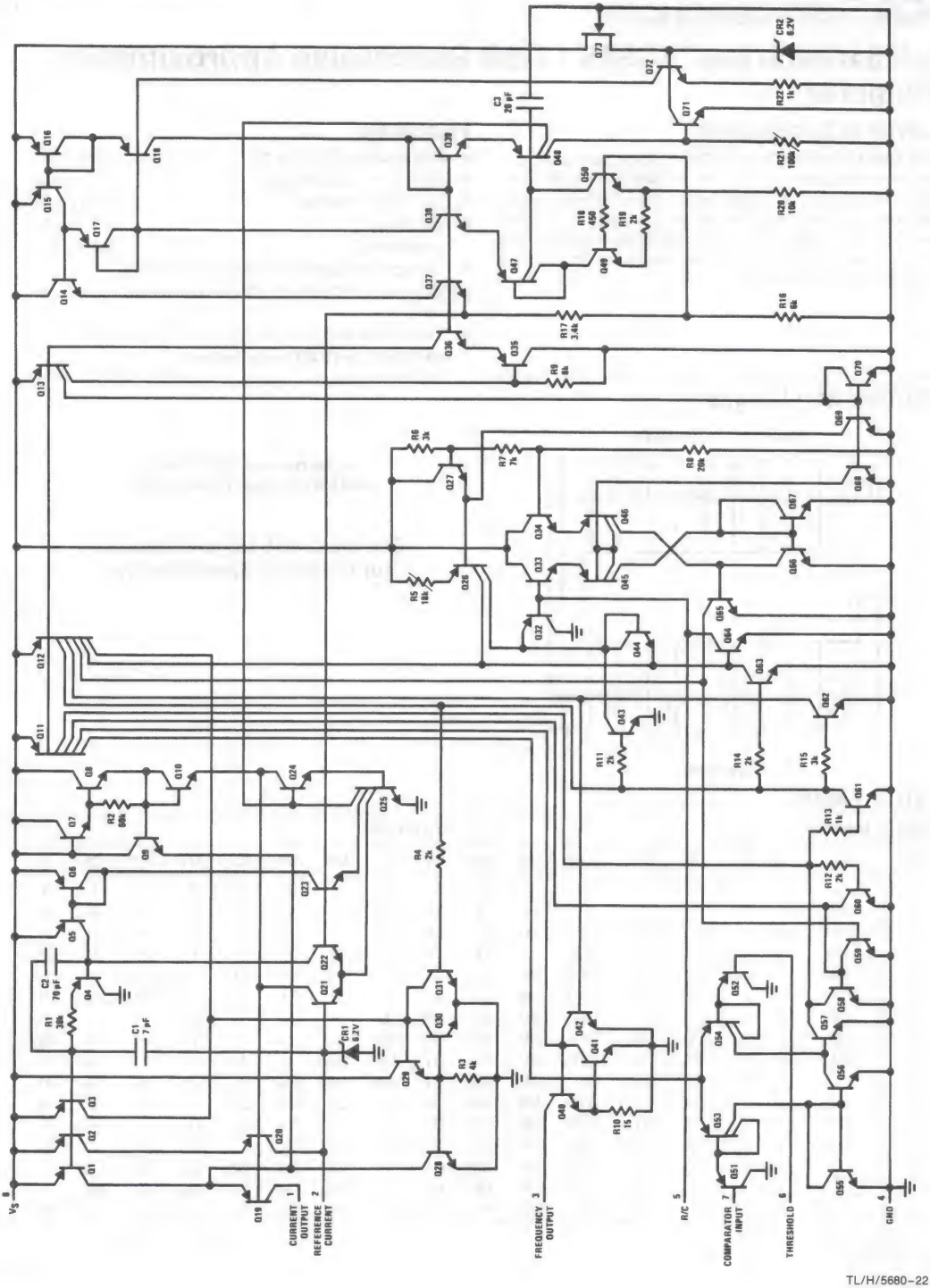
Dual-In-Line Package



TL/H/5680–21

Order Number LM231AN, LM231N, LM331AN, or LM331N
See NS Package Number N08E

Schematic Diagram



TL/H/5680–22

National Semiconductor

MM54C905/MM74C905 12-Bit Successive Approximation Register

General Description

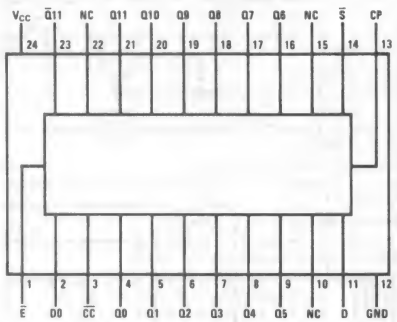
The MM54C905/MM74C905 CMOS 12-bit successive approximation register contains all the digit control and storage necessary for successive approximation analog-to-digital conversion. Because of the unique capability of CMOS to switch to each supply rail without any offset voltage, it can also be used in digital systems as the control and storage element in repetitive routines.

Features

- Wide supply voltage range 3.0V to15V
- Guaranteed noise margin 1.0V
- High noise immunity $0.45V_{CC}$ typ
- Low power TTL compatibility fan out of 2 driving 74L
- Provision for register extension or truncation
- Operates in START/STOP or continuous conversion mode
- Drive ladder switches directly. For 10 bits or less with 50k/100k R/2R ladder network

Connection Diagram

Dual-In-Line Package



TL/F/5712–1

Top View

Order Number MM74C905N
See NS Package Number N24C

See the CMOS Logic Databook for Complete Specifications

Truth Table

TIME	INPUTS			OUTPUTS													
t_n	D	$\overline{S}$	$\overline{E}$	D0	Q11	Q10	Q9	Q8	Q7	Q6	Q5	Q4	Q3	Q2	01	Q0	$\overline{C}$
0	X	L	L	X	X	X	X	X	X	X	X	X	X	X	X	X	X
1	D11	H	L	X	L	H	H	H	H	H	H	H	H	H	H	H	H
2	D10	H	L	D11	D11	L	H	H	H	H	H	H	H	H	H	H	H
3	D9	H	L	D10	D11	D10	L	H	H	H	H	H	H	H	H	H	H
4	D8	H	L	D9	D11	D10	D9	L	H	H	H	H	H	H	H	H	H
5	D7	H	L	D8	D11	D10	D9	D8	L	H	H	H	H	H	H	H	H
6	D6	H	L	D7	D11	D10	D9	D8	D7	L	H	H	H	H	H	H	H
7	D5	H	L	D6	D11	D10	D9	D8	D7	D6	L	H	H	H	H	H	H
8	D4	H	L	D5	D11	D10	D9	D8	D7	D6	D5	L	H	H	H	H	H
9	D3	H	L	D4	D11	D10	D9	D8	D7	D6	D5	D4	L	H	H	H	H
10	D2	H	L	D3	D11	D10	D9	D8	D7	D6	D5	D4	D3	L	H	H	H
11	D1	H	L	D2	D11	D10	D9	D8	D7	D6	D5	D4	D3	D2	L	H	H
12	D0	H	L	D1	D11	D10	D9	D8	D7	D6	D5	D4	D3	D2	D1	L	H
13	X	H	L	D0	D11	D10	D9	D8	D7	D6	D5	D4	D3	D2	D1	D0	L
14	X	X	L	X	D11	D10	D9	D8	D7	D6	D5	D4	D3	D2	D1	D0	L
	X	X	H	X	H	NC	NC	NC	NC	NC	NC	NC	NC	NC	NC	NC	NC

H = High level
L = Low level
X = Don't care
NC = No change

National Semiconductor

μA9708
6-Channel 8-Bit μP Compatible A/D Converter

General Description

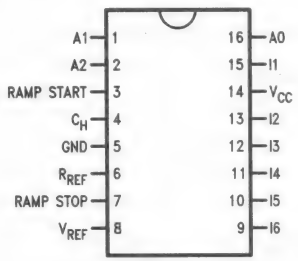
The μA9708 is a single slope 8-bit, 6-channel ADC subsystem that provides all of the necessary analog functions for a microprocessor-based data control system. The device uses an external microprocessor system to provide the necessary addressing, timing and counting functions and includes a 1-of-8 decoder, 8-channel analog multiplexer, sample and hold, ramp integrator, precision ramp reference, and a comparator on a single monolithic chip.

Features

- MPU compatible
- Excellent linearity over full temperature range ±0.2% maximum
- Typical 300 μs conversion time per channel
- Wide dynamic range includes ground
- Auto-zero and full-scale correction capability
- Ratiometric conversion—no precision reference required
- Single-supply operation
- TTL compatible
- Does not require access to data bus or address bus

Connection Diagram

All Packages



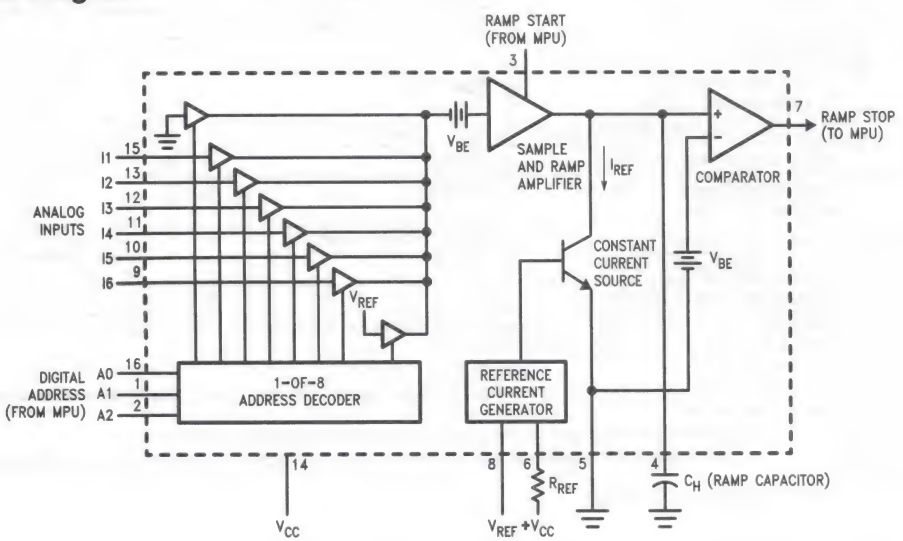
TL/H/10409–2

(Top View)

Order Number μA9708DC or μA9708DM
See NS Package Number J16A

Order Number μA9708PC
See NS Package Number N16E

Block Diagram



TL/H/10409–1

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	18V
Comparator Output (Ramp Stop)	−0.3V to +18V
Analog Input Range	−0.3V to +30V
Digital Input Range	−0.3V to +30V
Output Sink Current	10 mA
Storage Temperature Range	−65°C to +150°C
Continuous Total Dissipation	
Ceramic DIP Package	900 mW
Molded DIP Package	1000 mW
Pin Temperature	
Ceramic DIP (Soldering, 60 Sec.)	300°C
Molded DIP (Soldering, 10 Sec.)	260°C

Operating Ratings (Note 1)

Operating Temperature Range	
μA9708PC, μA9708DC	0°C to +70°C
μA9708DM	−55°C to +125°C
Supply Voltage (V_{CC})	4.75V to 15V
Reference Voltage (V_{REF}) (Note 2)	2.8V to 5.25V
Ramp Capacitor (C_H)	300 pF
Reference Current (I_R)	12 μA to 50 μA
Analog Input Range	0V to V_{REF}
Ramp Stop Output Current	1.6 mA

Electrical Characteristics

Over recommended operating conditions, V_{CC} = 5.0V, −55°C ≤ T_A ≤ +125°C for μA9708DM and 0°C ≤ T_A ≤ +70°C for μA9708DC or μA9708PC; unless otherwise specified.

Symbol	Parameter	Conditions	Min	Typ	Max	Units
E_A	Conversion Accuracy	Over Entire Temperature Range (Note 3)		±0.2	±0.3	%
E_R	Linearity	Applies to Any One Channel (Note 4)		±0.08	±0.2	%
V_{OSM}	Multiplexer Input Offset Voltage	Channel ON		2.0	4.0	mV
t_C	Conversion Time per Channel	Analog Input = 0V to V_{REF} C_H = 300 pF, I_{REF} = 50 μA		296	350	μs
t_A	Acquisition Time	C_H = 1000 pF		20	40	μs
I_A	Acquisition Current		150			μA
t_O	Ramp Start Delay Time			100		ns
t_M	Multiplexer Address Time			1.0		μs
V_{IH}	Digital Input HIGH Voltage	A0, A1, A2, Ramp Start	2.0			V
V_{IL}	Digital Input LOW Voltage	A0, A1, A2, Ramp Start			0.8	V
I_B	Analog Input Current	Channel ON or OFF	−3.0	−1.0		μA
I_{IL}	Input LOW Current	A0, A1, A2, Ramp Start = 0.4V	−15	−5		μA
I_{IH}	Input HIGH Current	A0, A1, A2, Ramp Start = 5.5V			1.0	μA
I_{OS}	Input Offset Current			1.0	3.0	μA
I_{OH}	Comparator Logic "1" Output Leakage Current	V_{OH} = 15V			10	μA
V_{OL}	Comparator Logic "0" Output Voltage	I_{OL} = 1.6 mA			0.4	V
PSRR	Power Supply Rejection Ratio	(Note 5)	40			dB
	Cross Talk between Any Two Channels	(Note 6)	60			dB
I_{CC}	Power Supply Current	V_{CC} = 5V to 15V, I0 = 0		7.5	15	mA
C_{IN}	Input Capacitance			3.0		pF
C_{OUT}	Comparator Output Capacitance			5.0		pF

Note 1: Absoute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but do not guarantee specific performance limits.

Note 2: V_{REF} should not exceed V_{CC} − 2V.

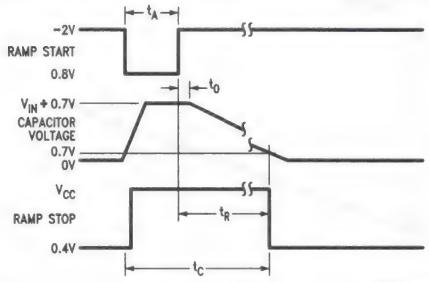
Note 3: Conversion accuracy is defined as the deviations from a straight line drawn between the points defined by channel address 000 (0 scale) and channel address 111 (full scale) for all channels.

Note 4: Linearity is defined as the deviation from a straight line drawn between the 0 and full scale points for each channel.

Note 5: Power supply rejection ratio is defined as the conversion error contributed by power supply voltage variations while resolving mid scale on any channel.

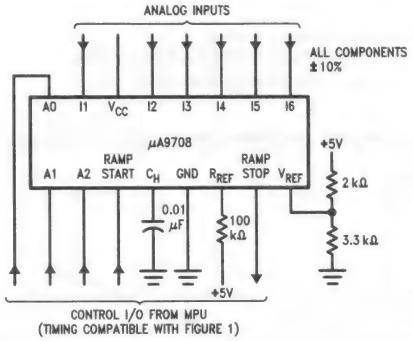
Note 6: Cross Talk between channels = $20 \log \frac{\Delta V_{CH}}{\Delta V_I}$.

Timing Diagram and Test Circuits



TL/H/10409–7

FIGURE 1. Equivalent Timing Waveform for Test Circuits and Applications



TL/H/10409–8

Input Timing:

$t_A > 400\ \mu s$

$$V_{REF} = \left(\frac{3.3\,k\Omega}{2\,k\Omega + 3.3\,k\Omega}\right) 5V = 3.1$$

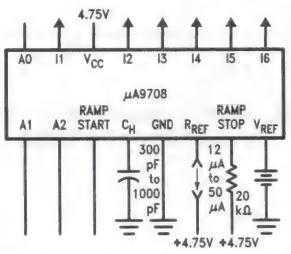
$$I_R = \frac{5 - 3.1}{100\,k\Omega} = 19\ \mu A$$

$t_R|_{max}$ = full scale ramp time

$$= \frac{0.01 \times 10^{-6}}{19 \times 10^{-6}} \times 3.1 = 1.6\ ms$$

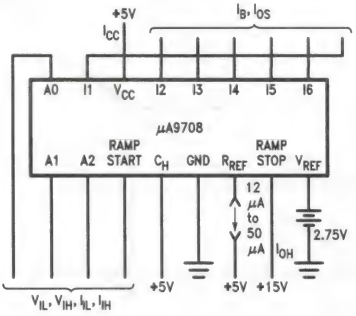
Note: For evaluation purposes,the ramp start timing generation can be implemented with an LM555 timer (astable operation) or MPU evaluation kit, and a time interval meter for ramp time measurement. The TIM meter will measure the time between to 0 to 1 transition of the ramp start and the 1 to 0 transition of the ramp stop. The ramp stop is open collector, and must have an external pull-up resistor to V_{CC}.

FIGURE 2. Slow Speed Evaluation Circuit for Ratiometric Operation



TL/H/10409–9

FIGURE 3. Linearity/Acquisition Time/ Conversion Time Test Circuit



TL/H/10409–10

FIGURE 4. Static Measurements

Functional Description

This Analog to Digital Converter is a single-slope 8-bit, 6-channel A/D converter that provides all of the necessary analog functions for a microprocessor-based data/control system. The device uses the processor system to provide the necessary addressing, timing and counting functions and includes a 1-of-8 decoder, 8-channel analog multiplexer, sample and hold, precision current reference, ramp integrator and comparator on a single monolithic chip.

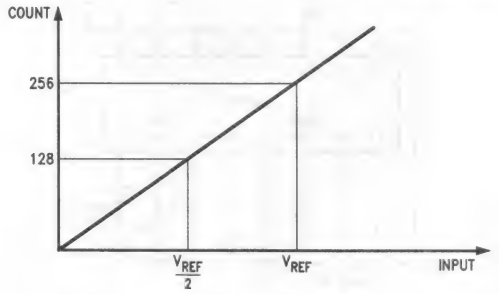
Applications that require auto-zero or auto-calibration, (See *Figures 5–8*) can use selection of address 000 and 111, for input address lines A0–A2, in conjunction with the arithmetic capability of a microprocessor to provide ground and scaling factors. Address 0, 0, 0 internally connects the input of the ramp generator to ground and may be used for zero offset correction in subsequent conversions. Address 1, 1, 1, internally connects the input of the ramp generator to the voltage reference, V_{REF}, and may be used for scale factor correction in subsequent conversions. For the following, refer to the Functional Block Diagram.

Six separate external analog voltage inputs may come into terminals I1–I6 and the specific analog input to be converted is selected via address terminals A0–A2. The analog input voltage level is transferred to the external ramp capacitor connected to pin 4 when the input to the ramp start terminal (pin 3) is at a logic 0 (See *Figure 1*). The time to charge the capacitor is the acquisition time which is a function of the output impedance of an amplifier internal to the A/D converter and the value of the capacitor. After charging the external capacitor the ramp start terminal is switched to a logic 1 which introduces a high impedance between the analog input voltage and the external capacitor.

The capacitor begins to discharge at a controlled rate. The controlled rate of discharge (ramp) is established by the external reference voltage, the external reference resistor, the value of the external capacitor and the internal leakage of the A/D converter. Connected to the capacitor terminal is a

Functional Description (Continued)

Auto-Zero and Full-Scale Features

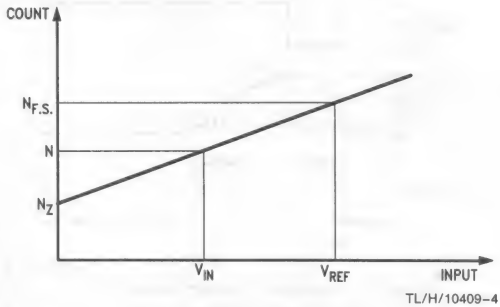


TL/H/10409–3

No Zero Offset
No Full-Scale Error

$$\text{Count (n)} = \frac{V_{IN}}{V_{REF}} \times 256$$

FIGURE 5. Ideal Transfer Function



TL/H/10409–4

$N_{F.S.} \neq 256$
$N_Z \neq 0$
(N) has both full-scale and zero errors

FIGURE 6. Transfer Function with Zero and Full-Scale Error

comparator internal to the A/D converter with its output going to the ramp stop terminal (pin 7). The comparator output is a logic one when the capacitor is charged and switches to a logic 0 when the capacitor is in a discharged state. The ramp time is from the time when ramp start goes HIGH (logic "1") to when ramp stop goes LOW (logic "0"). The microprocessor must be programmed to determine this conversion time. The ideal (no undesirable internal source impedances, leakage paths, errors on levels where comparator switches or delay time) conversion time is calculated as follows:

$$\text{Ramp Time} = V1 \frac{C_H}{I_R}$$

Where $V1$ = Analog Input Voltage Being Measured
C_H = External Ramp Capacitor

$$I_R = \frac{V_{CC} - V_{REF}}{R_{REF}}$$

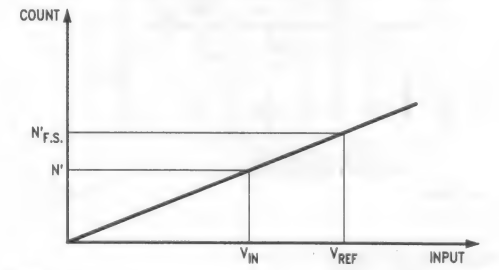
Where V_{CC} = Power Supply Voltage
V_{REF} = Reference Voltage
R_{REF} = Reference Resistor

In actual use the errors due to a nonideal A/D converter can be minimized by using a microprocessor to make the calculations. (See *Figures 5* through *8*.)

Channel Selection

Input Address Line			Selected Analog Input
A2	A1	A0	
0	0	0	Ground
0	0	1	I1
0	1	0	I2
0	1	1	I3
1	0	0	I4
1	0	1	I5
1	1	0	I6
1	1	1	V_{REF}

Auto-Zero and Full-Scale Features (Continued)

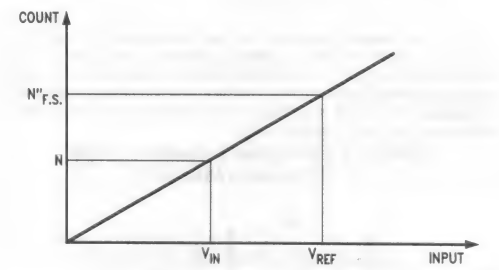


$N' = N - N_Z$

TL/H/10409–5

N' has Full-Scale Error

FIGURE 7. Transfer Functions with Zero-Correction Added



$$N'' = (N - N_Z) \times \frac{256}{(N_{F.S.} - N_Z)}$$

TL/H/10409–6

FIGURE 8. Transfer Function with both Zero and Full-Scale Correction Added

Typical Applications

Application Suggestions and Formulas

1. The capacitor node impedance is approximately 30 $\mu\Omega$ and should have no parallel resistance for proper operation.
2. t_R when $V_{IN} = 0V$ will be finite (i.e., the comparator will always toggle for $V_{IN} \geq 0V$).
3. The ramp stop output is open collector, and an external pull-up resistor is required.
4. All digital inputs and outputs are TTL compatible.
5. For proper operation, timing commences on the 0 to 1 transition of ramp start and terminates on the 1 to 0 transition of ramp stop.
6. $t_A \geq \dfrac{C_H}{150\ \mu A - I_R} \times V_{REF}$ (See *Figure 1*)
7. t_R (ramp time) $= \dfrac{C_H}{I_R \times V_{IN}}$, $t_R|_{max} = \dfrac{C_H}{I_R} \times V_{REF}$ (See *Figure 1*)
8. $I_R = \dfrac{V_{CC} - V_{REF}}{R_{REF}}$
9. $2V \leq V_{REF} \leq (V_{CC} - 2V)$
10. Address lines A0, A1, A2 must be stable throughout the sampling interval, t_A.
11. Pin 6 (R_{REF}) should be bypassed to ground via a 0.02 μF capacitor.

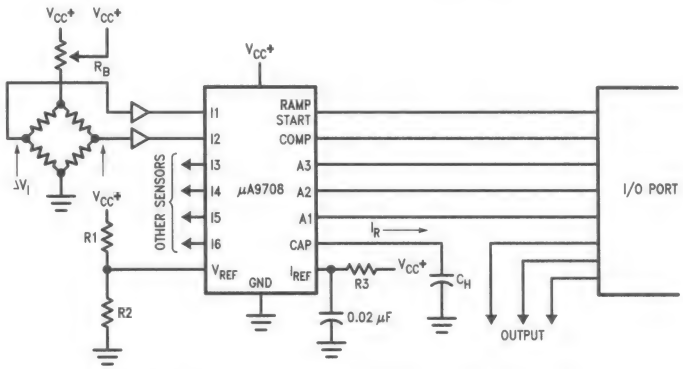
Microprocessor Considerations

Several alternatives exist from a hardware/software standpoint in microprocessor based systems using the μA9708.

1. The ramp time measurement may be implemented in software using a register increment, followed by a branch back depending on the status of the ramp stop.
2. Alternately, the ramp stop may be tied into the interrupt structure in systems containing a programmable binary timer. This scheme has the following advantages:
   a. The CPU is not committed during the ramp time interval.
   b. It requires only 4 bits of an I/O port for control signals.
3. The auto-zero/auto-full-scale (See *Figures 5–8*) should use double precision, rounded (as opposed to truncated) arithmatics. Several points are worth noting:
   a. The subtractions are single op code instructions.
   b. The full scale correction uses a multiply by 256 and can be accomplished by a shift left 8 bits (usually one instruction) or placing ($N - N_Z$) in the MSB register and setting the LSB register to zero, for the double precision divide.
   c. The divisor ($N_{F.S.} - N_Z$) of the MSB register will always be zero.

These schemes have the following advantages:

a. No access to the data bus or address bus is required, by the A/D system.
b. 4 I/O bits completely support the A/D system.
c. Since auto full scale/auto zero are implemented in software and long term drift (aging) effects are eliminated.
d. Software overhead is minimal (typically 30 bytes).
e. Where ratiometric operation is permissible, the 4 external components may be ±5% tolerance, including the power supply.

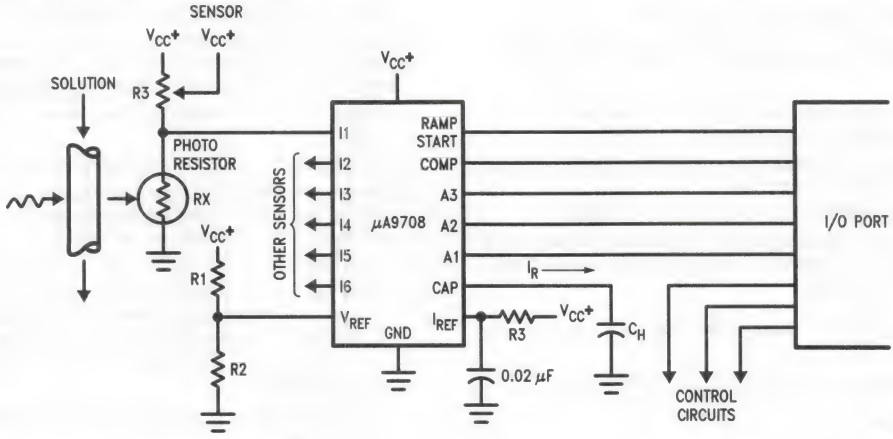


TL/H/10409–11

Note: ΔV_I = (Applied Force) and can be Linearized (if necessary) in Software.

FIGURE 9. Ratiometric Strain Gage Sensore/Controller

Typical Applications (Continued)



TL/H/10409–12

Applications
Beverage Brewers/Dispensers
Chemical Solution Control
Automatic Liquid Mixing Control

$$\text{Ramp Current} = I_R = V_{CC}\left(\frac{R_1}{R_1 + R_2}\right)\left(\frac{1}{R_3}\right)$$

$$V_I = \left(\frac{R_X}{R_X + R_B}\right) V_{CC+}$$

$$\text{Ramp Time} = V_I\left(\frac{C_H}{I_R}\right) = \left(\frac{R_X}{R_X + R_B}\right)\left(1 + \frac{R_2}{R_1}\right)\left(C_H R_3\right)$$

FIGURE 10

Section 4

Digital-to-Analog Converters

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Definition of Terms
D/A Converters

Conversion Time: The time required for a complete measurement by an analog-to-digital converter.

DC Common-Mode Error: This specification applies to ADCs with differential inputs. It is the change in the output code that occurs when the analog voltages on the two inputs are changed by an equal amount. It is expressed in LSBs.

Differential Nonlinearity: Ideally, any two adjacent digital codes correspond to measured analog voltages that are exactly one LSB apart. Differential non-linearity is a measure of the worst case deviation from the ideal 1 LSB step. For example, a DAC with a 1.5 LSB output change for a 1 LSB digital code change exhibits ½ LSB differential non-linearity. Differential non-linearity may be expressed in fractional bits or as a percentage of full scale. A differential non-linearity greater than 1 LSB will lead to a non-monotonic transfer function in a DAC and missing codes in an ADC.

Gain Error (Full Scale Error): For an ADC, the difference (usually expressed in LSBs) between the input voltage that should ideally produce a full scale output code and the actual input voltage that produces that code. For DACs, it is the difference between the output voltage (or current) with full scale input code and the ideal voltage (or current) that should exist with a full scale input code.

Gain Temperature Coefficient (Full Scale Temperature Coefficient): Change in gain error divided by change in temperature. Usually expressed in parts per million per degree Celsius (ppm/°C).

Integral Nonlinearity (Linearity Error): Worst case deviation from the line between the endpoints (zero and full scale). Can be expressed as a percentage of full scale or in fractions of an LSB.

LSB (Least-Significant Bit): In a binary coded system this is the bit that carries the smallest value or weight. Its value is the full scale voltage (or current) divided by 2^n, where n is the resolution of the converter.

Monotonicity: A monotonic function has a slope whose sign does not change. A monotonic DAC has an output that changes in the same direction (or remains constant) for each increase in the input code. The converse is true for decreasing codes.

MSB (Most Significant Bit): In a binary coded system this is the bit that has the largest value or weight. Its value is one half of full scale.

Multiplying DAC: In a sense, every DAC is a multiplying DAC since the output voltage (or current) is equal to the reference voltage times a constant determined by the digital input code divided by 2^n (n is the number of bits of resolution). In a two quadrant multiplying DAC the reference voltage or the digital input code can change the output voltage polarity. If both the reference voltage and the digital code change the output voltage polarity, four quadrant multiplication exists.

Offset Error (Zero Error): In a DAC, this is the output voltage that exists when the input digital code is set to give an ideal output of zero volts. In the case of an ADC, this is the difference between the ideal input voltage (½ LSB) and the actual input voltage that is needed to make the transition from zero to 1 LSB. All the digital codes in the transfer curve are offset by the same value. Many converters allow nulling of offset with an external potentiometer. Offset error is usually expressed in LSBs.

Power Supply Rejection (Power Supply Sensitivity): The sensitivity of a converter to changes in the dc power supply voltages.

Quantizing Error: The error inherent in all A/D conversions. Since even an "ideal" converter has finite resolution, any analog voltage that falls between two adjacent output codes will result in an output code that is inaccurate by up to ½ LSB.

Ratiometric Operation: Many A/D applications require a stable and accurate reference voltage against which the input voltage is compared. This approach results in an **absolute conversion**. Some applications, however, use transducers or other signal sources whose output voltages are proportional to some external reference. In these **ratiometric** applications, the reference for the signal source should be connected to the reference input of the converter. Thus, any variations in the source reference voltage will also change the converter reference voltage and produce an accurate conversion.

Resolution: The smallest analog increment corresponding to a 1 LSB converter code change. For converters, resolution is normally expressed in bits, where the number of analog levels is equal to 2^n. As an example, a 12-bit converter divides the analog signal into $2^{12} = 4096$ discrete voltage (or current) levels.

Settling Time: The time from a change in input code until a DAC's output signal remains within ± ½ LSB (or some other specified tolerance) of the final value.

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D/A Converter Selection Guide

Part No.	Resolution (Bits)	Linearity @ 25°C % (Max)	Settling Time (+½ LSB)	Supplies (V)	Temperature Range*			Package	Comments
					M	I	C		
DAC0631	6	0.78	28	5			•	28-Pin DIP 44-Pin PCC	Triple 35 MHz Video DAC
DAC0631-40	6	0.78	25	5			•	28-Pin DIP 44-Pin PCC	Triple 40 MHz Video DAC
DAC0630	6	0.78	20	5			•	28-Pin DIP 44-Pin PCC	Triple 50 MHz Video DAC
ADC0852	8	0.19		5		•	•	8-Pin DIP	DAC, Comparator, Serial Input
ADC0854	8	0.19		5		•	•	14-Pin DIP	DAC, Comparator, Serial Input
DAC0800	8	0.19	100 ns	±5 to ±15	•		•	16-Pin DIP 16-Pin S.O.	High-Speed Multiplying
DAC0801	8	0.39	100 ns	±5 to ±15	•		•	16-Pin DIP 16-Pin S.O.	High-Speed Multiplying
DAC0802	8	0.10	100 ns	±5 to ±15	•		•	16-Pin DIP 16-Pin S.O.	High-Speed Multiplying
DAC0806	8	0.78	150 ns	±5 to ±15			•	16-Pin DIP 16-Pin S.O.	Multiplying
DAC0807	8	0.39	150 ns	±5 to ±15			•	16-Pin DIP 16-Pin S.O.	Multiplying
DAC0808	8	0.19	150 ns	±5 to ±15	•		•	16-Pin DIP 16-Pin S.O.	Multiplying
DAC0830	8	0.05	1 μs	5 to 15	•	•	•	20-Pin DIP 20-Pin S.O. 20-Pin PCC	μP Compatible 4-Quadrant Multiplying
DAC0831	8	0.10	1 μs	5 to 15			•	20-Pin DIP	μP Compatible 4-Quadrant Multiplying
DAC0832	8	0.20	1 μs	5 to 15		•	•	20-Pin DIP 20-Pin S.O. 20-Pin PCC	μP Compatible 4-Quadrant Multiplying
DAC1000	10	0.05	500 ns	5 to 15	•	•	•	24-Pin DIP	μP Compatible Double Buffered
DAC1001	10	0.1	500 ns	5 to 15			•	24-Pin DIP	μP Compatible Double Buffered
DAC1002	10	0.2	500 ns	5 to 15	•	•	•	24-Pin DIP	μP Compatible Double Buffered
DAC1006	10	0.05	500 ns	5 to 15	•	•	•	20-Pin DIP	μP Compatible Double Buffered
DAC1007	10	0.1	500 ns	5 to 15		•	•	20-Pin DIP	μP Compatible Double Buffered
DAC1008	10	0.2	500 ns	5 to 15	•	•	•	20-Pin DIP	μP Compatible Double Buffered

D/A Converter Selection Guide (Continued)

Part No.	Resolution (Bits)	Linearity @ 25°C % (Max)	Settling Time (+½ LSB)	Supplies (V)	Temperature Range*			Package	Comments
					M	I	C		
DAC1020	10	0.05	500 ns	5 to 15	•	•	•	16-Pin DIP	4-Quadrant Multiplying
DAC1021	10	0.1	500 ns	5 to 15	•	•	•	16-Pin DIP	4-Quadrant Multiplying
DAC1022	10	0.2	500 ns	5 to 15	•	•	•	16-Pin DIP	4-Quadrant Multiplying
DAC1208	12	0.012	1 μs	5 to 15		•	•	24-Pin DIP	μP Compatible 4-Quadrant Multiplying
DAC1209	12	0.024	1 μs	5 to 15		•	•	24-Pin DIP	μP Compatible 4-Quadrant Multiplying
DAC1210	12	0.05	1 μs	5 to 15		•	•	24-Pin DIP	μP Compatible 4-Quadrant Multiplying
DAC1218	12	0.012	1 μs	5 to 15		•	•	18-Pin DIP	4-Quadrant Multiplying
DAC1219	12	0.024	1 μs	5 to 15		•	•	18-Pin DIP	4-Quadrant Multiplying
DAC1220	12	0.05	500 ns	5 to 15	•	•	•	18-Pin DIP	4-Quadrant Multiplying
DAC1221	12	0.1	500 ns	5 to 15			•	18-Pin DIP	4-Quadrant Multiplying
DAC1222	12	0.2	500 ns	5 to 15	•	•	•	18-Pin DIP	4-Quadrant Multiplying
DAC1230	12	0.012	1 μs	5 to 15		•	•	20-Pin DIP	μP Compatible 4-Quadrant Multiplying
DAC1231	12	0.024	1 μs	5 to 15		•	•	20-Pin DIP	μP Compatible 4-Quadrant Multiplying
DAC1232	12	0.05	1 μs	5 to 15		•	•	20-Pin DIP	μP Compatible 4-Quadrant Multiplying
DAC1265A	12	0.006	200 ns	±15	•		•	24-Pin DIP	High-Speed
DAC1265	12	0.012	200 ns	±15	•		•	24-Pin DIP	High-Speed
DAC1266A	12	0.006	200 ns	±12 to ±15	•		•	24-Pin DIP	High-Speed
DAC1266	12	0.012	200 ns	±12 to ±15	•		•	24-Pin DIP	High-Speed

*Ambient temperature range for "M" is −55°C to +125°C, "I" is −25°C to +85°C or −40°C to +85°C, "C" 0°C to +70°C.

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DAC0630/DAC0631
Triple 6-Bit Video DAC with Color Palette

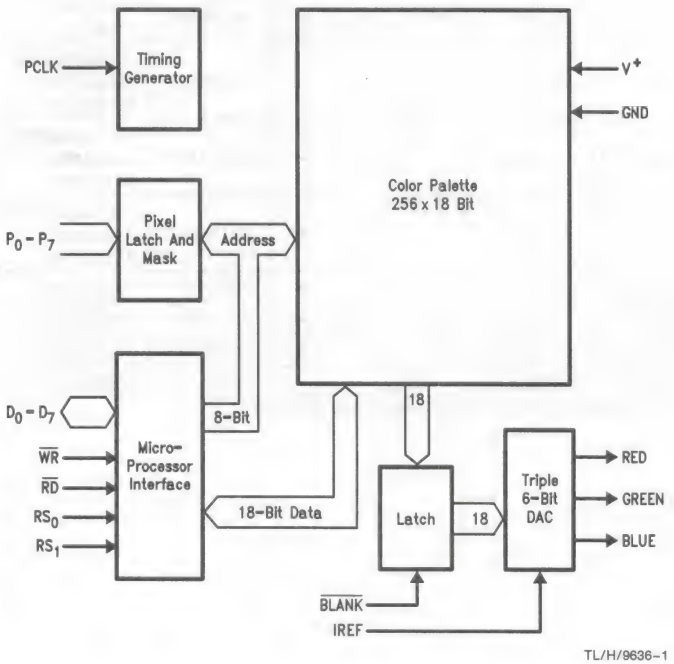
General Description

The DAC0630 and DAC0631 are monolithic triple 6-bit video digital-to-analog converters with on-chip 256 x 18 bit color palettes and are intended for graphics applications. The color palette makes possible the display of 256 colors selected from a total of 256K possible colors through the internal 6-bit video DACs. The DACs are capable of driving 75Ω or 37.5Ω loads to normal video levels at pixel rates of 50 MHz (DAC0630) and 35 MHz (DAC0631). The DAC0630 and DAC0631 provide a bi-directional microprocessor interface with TTL compatible inputs. The DAC0630 and DAC0631 are pin- and functionally-compatible with the Inmos IMS G171-50 and IMS G171-35 and IMS G176-50 and IMS G176-35.

Features

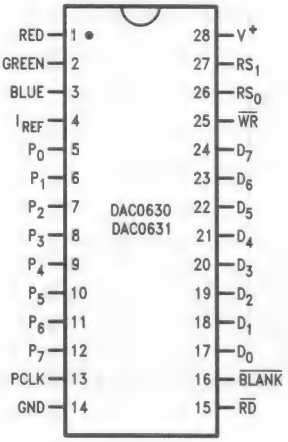
- Pixel rates of 50 MHz (DAC0630) and 35 MHz (DAC0631)
- 256 x 18 bit color palette
- 256K possible colors
- Color palette read-back
- Three internal 6-bit DACs
- Directly drives (75Ω) video cable
- RGB analog output
- Composite blank
- Single +5V supply
- Low power, high performance CMOS/bipolar processing
- TTL compatible inputs
- Full asynchronous μP interface
- 28-pin package

Block and Connection Diagrams



TL/H/9636–1

Dual-In-Line Package



TL/H/9636–2

Top View

Order Number
DAC0630CCD or DAC0631CCD
See NS Package Number D28D

Order Number DAC0631CCN
See NS Package Number N28B

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Positive Supply Voltage (V^+)	GND − 0.3V to 7V
Voltage at Logic Inputs (Note 3)	GND − 0.5V to V^+ + 0.5V
Voltage at Analog Pins 1–4 (Note 3)	GND − 0.5V to V^+ + 0.5V
Analog Output Current, Pins 1–3	45 mA
Reference Current, Pin 4	15 mA
DC Digital Output Current (Note 4)	25 mA
Power Dissipation (Note 5)	1.0W
ESD Susceptability (Note 6)	2000V
Soldering Information	
D Package (10 sec)	300°C
N Package (10 sec)	260°C
Storage Temperature	−65°C to 150°C

Operating Ratings (Notes 1 & 2)

Temperature Range	
$T_{MIN} \le T_A \le T_{MAX}$	$0°C \le T_A \le +70°C$
Positive Supply Voltage	4.5 to 5.5V

AC and DC Electrical Characteristics

The following specifications apply for $V^+ = +5V$, unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = 25°C$.

Symbol	Parameter		Conditions	DAC0630 DAC0631 Typical (Note 7)	DAC0630 DAC0631 Tested Limit (Note 8)	DAC0630 DAC0631 Design Limit (Note 9)	Units
I_{REF}	Reference Current	Minimum			**−3**		mA
		Maximum			**−10**		mA
I_{AVE}	Maximum Average Supply Current	DAC0630	P_{CLK} = 50 MHz I_{REF} = 10 mA Digital Outputs Unloaded		**160**		mA
		DAC0631	P_{CLK} = 35 MHz I_{REF} = 10 mA Digital Outputs Unloaded		**150**		mA
V_{REFmin}	Minimum Reference Voltage at I_{REF} Pin		$V^+ = 4.5V$ I_{REF} = 8.88 mA		**$V^+ - 3$**		V
I_{IN}	Maximum Digital Input Current (Pins 5–13, 15, 16, 25–27)		$V^+ = 5.5V$ $GND \le V_{IN} \le V^+$		**±10**		μA
I_{OZ}	Maximum Tri-State Digital Output Current (Pins 17–24)		$V^+ = 5.5V$ $GND \le V_{IN} \le V^+$		**±50**		μA
V_{OH}	Minimum Logic "1" Output Voltage		$V^+ = 4.5V$, $I_O = -5$ mA		**2.4**		V
V_{OL}	Maximum Logic "0" Output Voltage		$V^+ = 4.5V$, $I_O = +5$ mA		**0.4**		V
V_{IH}	Minimum Logic "1" Input Voltage		$4.5V \le V^+ \le 5.5V$		**2**		V
V_{IL}	Maximum Logic "0" Input Voltage		$4.5V \le V^+ \le 5.5V$		**0.8**		V
	DAC Resolution				**6**		Bits
V_{OUT}	Minimum Output Voltage Compliance (Pins 1–3)		$I_{OUT} \le 10$ mA		**1.5**		V

AC and DC Electrical Characteristics (Continued)

The following specifications apply for $V^+ = +5V$, unless otherwise specified. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = 25°C$.

Symbol	Parameter		Conditions	DAC0630 DAC0631			Units
				Typical (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	
I_{OUT}	Maximum Output Current Compliance (Pins 1–3)		$V_{OUT} \leq 1V$ $I_{REF} \leq 10$ mA		**21**		mA
	Full-Scale Gain Error (Note 10)		$Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA $Z_L = 37.5\Omega + 30$ pF $I_{REF} = 8.88$ mA		$-8, +2$ $-14, -4$		% %
	DAC-to-DAC Mismatch		$Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA (See Note 11)		**±2**		%
	Integral Non-Linearity (Note 12)		$Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA		**±0.5**		LSB
t_{ON}	Rise Time (Note 13)		$Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA			8	ns
	Maximum Full-Scale Settling Time	DAC0630 DAC0631	$Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA (See Note 14)			20 28	ns ns
	Maximum Glitch Energy		$Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA (See Note 15)	±200		±400	pV-sec
C_{IN}	Digital Input Capacitance (Pins 5–13, 15, 16, 25–27)			7			pF
C_{OUT}	Digital Output Capacitance (Pins 17–24)		$\overline{RD}$ = Logic High	7			pF
C_{OUTA}	Analog Output Capacitance (Pins 1–3)		$\overline{BLANK}$ = Logic Low	10			pF
$V_{OUTBLANK}$	Maximum Blanking Output Voltage		$\overline{BLANK}$ = Logic Low $Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA		**±0.5**		LSB
	Unadjusted Output Offset Error		$\overline{BLANK}$ = Logic High $Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA		**±0.5**		LSB
	Clock Feedthrough (Note 16)	DAC0630D DAC0631D DAC0631N	$P_{CLK} = 50$ MHz $P_{CLK} = 35$ MHz $P_{CLK} = 35$ MHz $Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA			−30 −35 −30	dB dB dB
PSS	Power Supply Sensitivity		$4.5V \leq V^+ \leq 5.5V$ I_{OUT} = Full Scale $Z_L = 75\Omega + 30$ pF $I_{REF} = 4.44$ mA		**6**		%/V

AC Electrical Characteristics

The following specifications apply for $V^+ = +5V$. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = 25°C$. Design Limits apply for $4.5V \le V^+ \le 5.5V$.

Symbol	Parameter		Conditions	DAC0630			DAC0631			Units
				Typical (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Typical (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	
t_{CHCH}	Minimum PCLK Period				20	**20**		28	**28**	ns
Δt_{CHCH}	Maximum PCLK Jitter		(Note 17)	±2.5			±2.5			%
t_{CLCH}	Minimum PCLK Width Low				6	**6**		9	**9**	ns
t_{CHCL}	Minimum PCLK Width High				6	**6**		7	**7**	ns
t_{PVCH}	Minimum Pixel Word Setup Time		(Note 18)		4	**4**		4	**4**	ns
t_{CHPX}	Minimum Pixel Word Hold Time		(Note 18)		4	**4**		4	**4**	ns
t_{BVCH}	Minimum $\overline{BLANK}$ Setup Time		(Note 18)		4	**4**		4	**4**	ns
t_{CHBX}	Minimum $\overline{BLANK}$ Hold Time		(Note 18)		4	**4**		4	**4**	ns
t_{CHAV}	PCLK to Valid DAC Output	Minimum	(Note 19)		5	**5**		5	**5**	ns
		Maximum			30	**30**		30	**30**	
Δt_{CHAV}	Maximum Differential Output Delay		(Note 20)	1			1			ns
t_{WLWH}	Minimum $\overline{WR}$ Pulse Width Low				50	**50**		50	**50**	ns
t_{RLRH}	Minimum $\overline{RD}$ Pulse Width Low				50	**50**		50	**50**	ns
t_{SVWL}	Minimum Register Select Setup Time		(Write Cycle)		10	**10**		15	**15**	ns
t_{SVRL}	Minimum Register Select Setup Time		(Read Cycle)		10	**10**		15	**15**	ns
t_{WLSX}	Minimum Register Select Hold Time		(Write Cycle)		10	**10**		15	**15**	ns
t_{RLSX}	Minimum Register Select Hold Time		(Read Cycle)		10	**10**		15	**15**	ns
t_{DVWH}	Minimum $\overline{WR}$ Data Setup Time				10	**10**		15	**15**	ns
t_{WHDX}	Minimum $\overline{WR}$ Data Hold Time				10	**10**		15	**15**	ns
t_{RLQX}	Minimum Output Turn-On Delay				5	**5**		5	**5**	ns
t_{RLQV}	Maximum $\overline{RD}$ Enable Access Time				40	**40**		40	**40**	ns
t_{RHQX}	Minimum Output Hold Time				5	**5**		5	**5**	ns
t_{RHQZ}	Maximum Output Turn-Off Delay		(Note 21)		20	**20**		20	**20**	ns
t_{WHWL1}	Minimum Successive Write Interval				$3(t_{CHCH})$	**$3(t_{CHCH})$**		$3(t_{CHCH})$	**$3(t_{CHCH})$**	
t_{WHRL1}	Minimum $\overline{WR}$ followed by Read Interval				$3(t_{CHCH})$	**$3(t_{CHCH})$**		$3(t_{CHCH})$	**$3(t_{CHCH})$**	
t_{RHRL1}	Minimum Successive Read Interval				$3(t_{CHCH})$	**$3(t_{CHCH})$**		$3(t_{CHCH})$	**$3(t_{CHCH})$**	
t_{RHWL1}	Minimum $\overline{RD}$ followed by Write Interval				$3(t_{CHCH})$	**$3(t_{CHCH})$**		$3(t_{CHCH})$	**$3(t_{CHCH})$**	
t_{WHWL2}	Minimum $\overline{WR}$ after Color Write		(Note 22)		$3(t_{CHCH})$	**$3(t_{CHCH})$**		$3(t_{CHCH})$	**$3(t_{CHCH})$**	

AC Electrical Characteristics (Continued) The following specifications apply for $V^+ = +5V$. **Boldface limits apply for T_{MIN} to T_{MAX};** all other limits $T_A = 25°C$. Design Limits apply for $4.5V \le V^+ \le 5.5V$.

Symbol	Parameter	Conditions	DAC0630			DAC0631			Units
			Typical (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	Typical (Note 7)	Tested Limit (Note 8)	Design Limit (Note 9)	
t_{WHRL2}	Minimum $\overline{RD}$ after Color Write	(Note 22)		$3(t_{CHCH})$	**$3(t_{CHCH})$**		$3(t_{CHCH})$	**$3(t_{CHCH})$**	
t_{RHRL2}	Minimum $\overline{RD}$ after Color Read	(Note 22)		$6(t_{CHCH})$	**$6(t_{CHCH})$**		$6(t_{CHCH})$	**$6(t_{CHCH})$**	
t_{RHWL2}	Minimum $\overline{WR}$ after Color Read	(Note 22)		$6(t_{CHCH})$	**$6(t_{CHCH})$**		$6(t_{CHCH})$	**$6(t_{CHCH})$**	
t_{WHRL3}	Minimum $\overline{RD}$ after Read Address Write	(Note 22)		$6(t_{CHCH})$	**$6(t_{CHCH})$**		$6(t_{CHCH})$	**$6(t_{CHCH})$**	
	Maximum Write/Read Enable Transition Time				**50**			**50**	ns

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating ratings.

Note 2: All voltages are measured with respect to ground, unless otherwise specified.

Note 3: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < GND$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 5 mA or less. The sum of the currents at all pins that are driven beyond the power supply voltages should not exceed 20 mA.

Note 4: One output at any time. The maximum time for this output level is one second.

Note 5: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 125°C$, and the typical thermal resistance (θ_{JA}) of the DAC0630/0631CCD when board mounted is 40°C/W. The typical thermal resistance for the DAC0630/631CCN when board mounted is 85°C/W.

Note 6: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 7: Typicals are at 25°C and represent most likely parametric norm.

Note 8: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 9: Design limits are guaranteed to National's AOQL (Average Outgoing Quality Level) but not 100% tested.

Note 10: Full-Scale Gain Error is defined as $\{[(F.S.\ I_{OUT})R_L - 2.1(I_{REF})R_L]/[2.1(I_{REF})R_L]\}100\%$. $V_{BLACK\ LEVEL} = 0V$.

Note 11: The listed value is relative to the midpoint of the full-scale distribution of the internal three DACs.

Note 12: Zero and full-scale adjusted linearity error $= [V_{out} - V_{offset} - (D \times V_{LSB})]/V_{LSB}$, $V_{LSB} =$ (Vfull scale−Voffset)/63.

Note 13: The rise time is measured from 10% to 90% of the full scale transition.

Note 14: The output signal's settling time is measured from a 2% change at the transition's initial value until it has settled to within 2% of the final value, excluding clock feedthrough.

Note 15: This value is determined using triangle approximation: glitch energy = (area of positive transient)−(area of negative transient).

Note 16: The value shown is the ratio of the RMS value of any PCLK signal on the analog outputs to the full-scale output voltage (700 mV).

Note 17: This parameter is the allowed variation in the pixel clock frequency. It does not permit the pixel clock period to vary below the minimum value for pixel clock (t_{CHCH}) period specified above.

Note 18: It is necessary that the color palette's pixel address be a valid logic level with the appropriate setup and hold times at each rising edge of P_{CLK} (this requirement includes the blanking period).

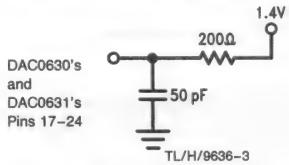
Note 19: A valid analog output is defined as the 50% point between successive values. This parameter is stable with time but can vary between different devices and may vary with different dc operating conditions.

Note 20: This applies to different analog outputs on the same device.

Note 21: Measured at ±200 mV from initial steady state output voltage.

Note 22: This parameter allows synchronization between operations on the microprocessor interface and the pixel stream being processed by the color palette.

AC Test Conditions Digital Output Load



Input Pulse Levels ..... GND to 3V
Input Rise and Fall Times (10% to 90%) ..... 2.5 ns
Digital Input Timing Reference Level 1.5V
Digital Output Timing Reference Level ..... 0.8V and 2.4V

Timing Waveforms

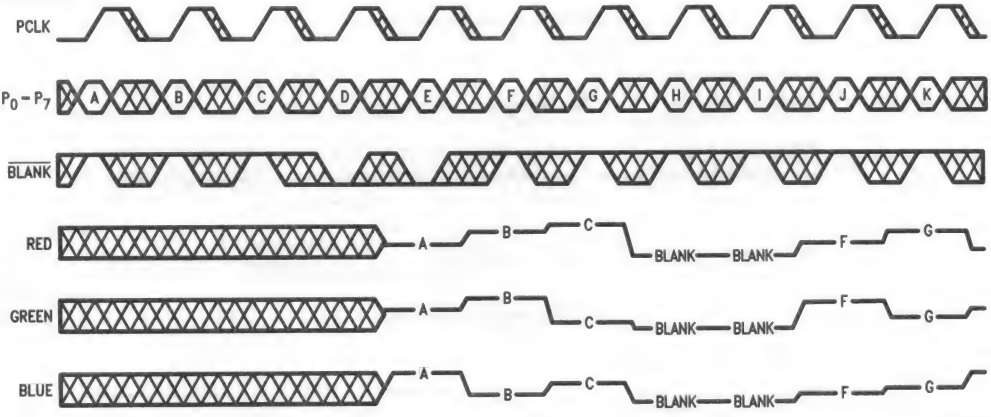


FIGURE 1. System Timing Diagram

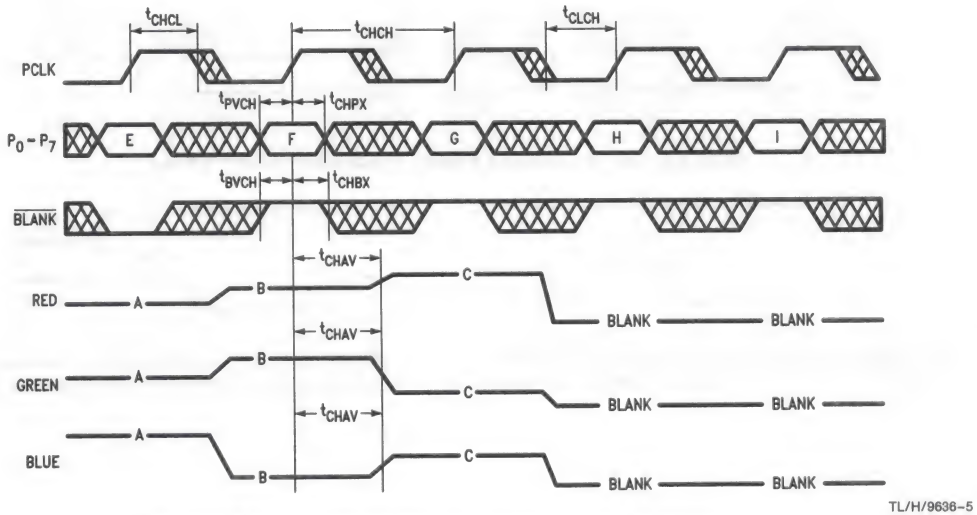


FIGURE 2. Expanded Timing Diagram Detailing Timing Specifications

Timing Waveforms (Continued)

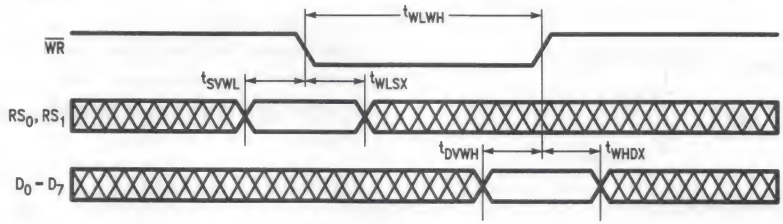


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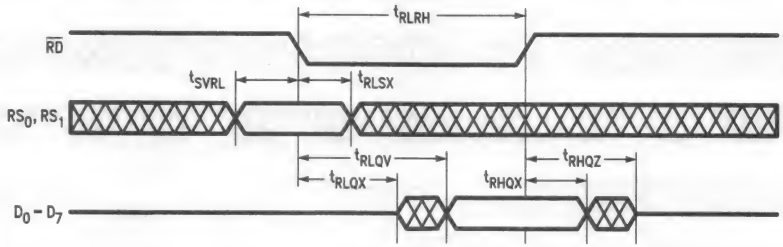
FIGURE 3. Basic Write Cycle



TL/H/9636–7

FIGURE 4. Basic Read Cycle

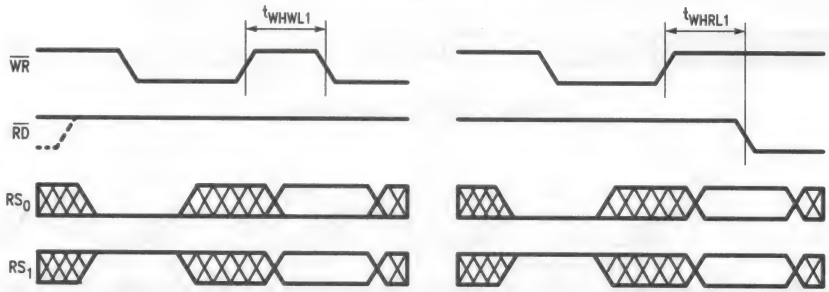


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FIGURE 5. Write to Pixel Mask Register Followed by a)Write, b)Read

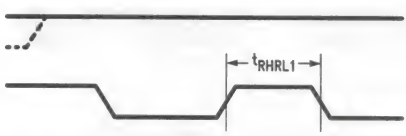

FIGURE 6a. Read from Pixel Mask or Pixel Address Register (Read or Write Mode) Followed by Read

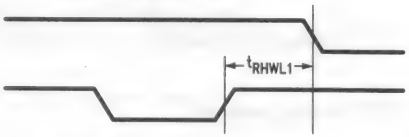

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FIGURE 6b. Read from Pixel Mask or Pixel Address Register (Read or Write Mode) Followed by Write

Timing Waveforms (Continued)

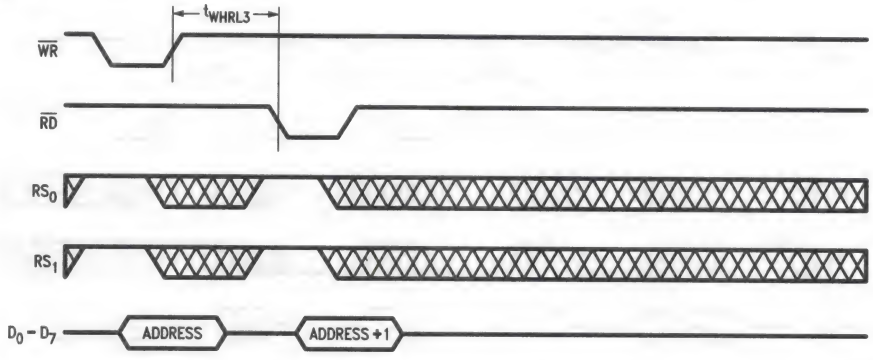

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FIGURE 7. Write and Read Back Pixel Address Register (Read Mode)

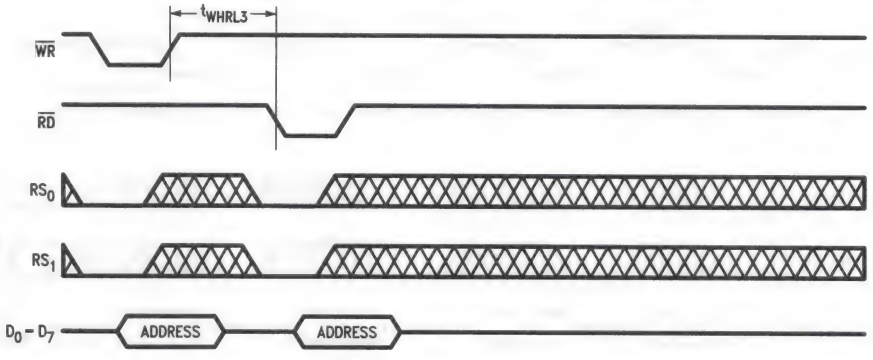

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FIGURE 8. Write and Read Back Pixel Address Register (Write Mode)

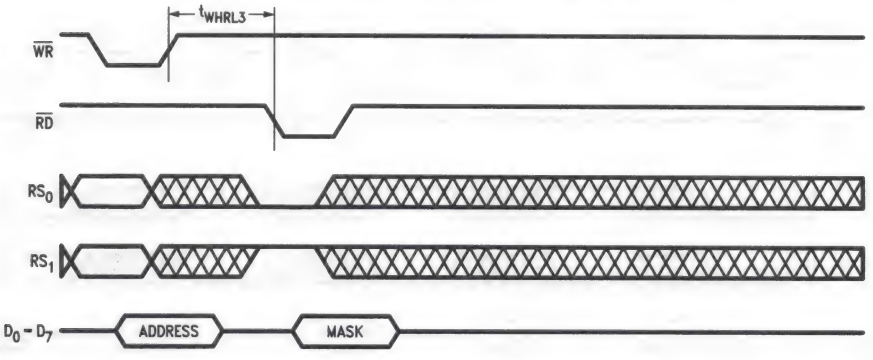

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FIGURE 9. Write Pixel Address Register (Read or Write Mode) then Read Pixel Mask Register

Timing Waveforms (Continued)

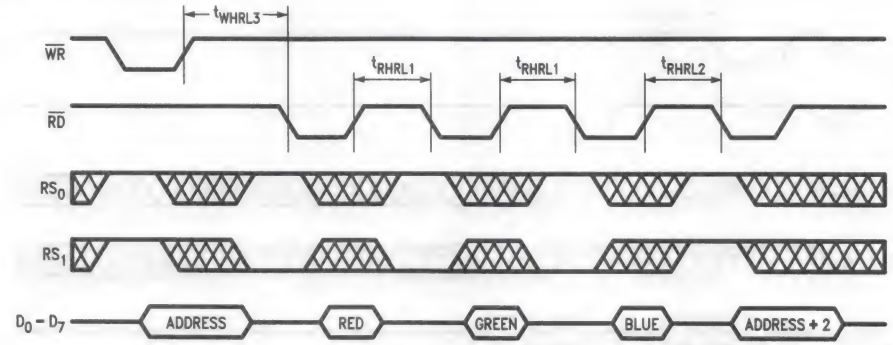

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FIGURE 10. Read Color Value then Read Pixel Address Register (Read Mode)

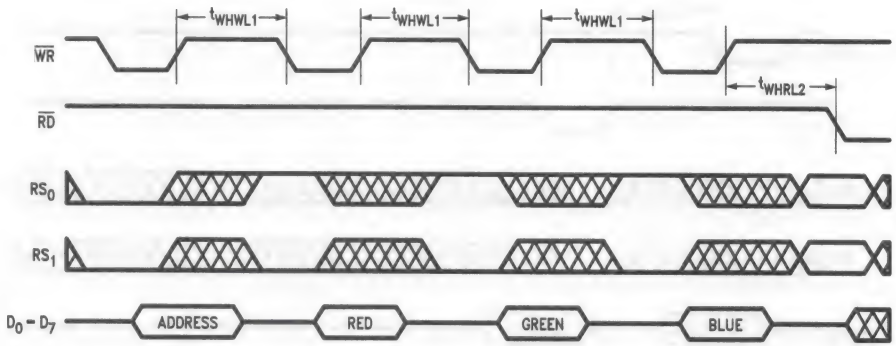

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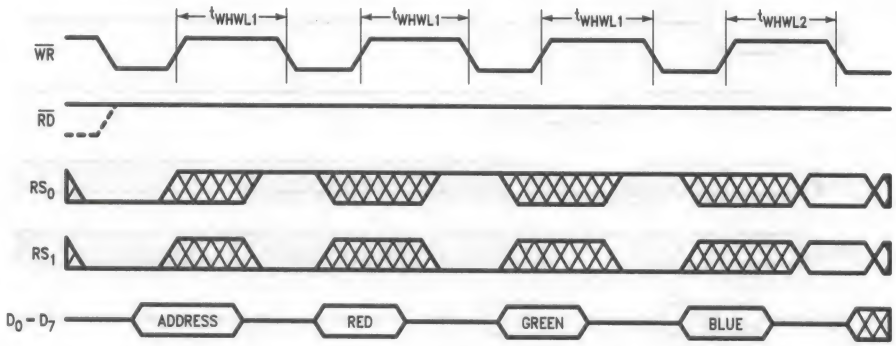
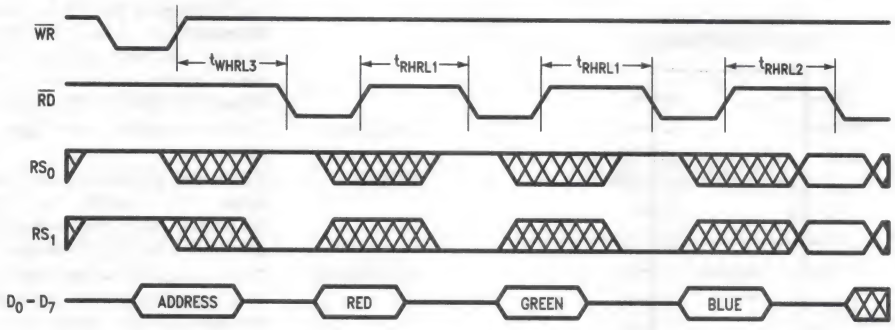
FIGURE 11. Color Value Write Followed by Any Read



TL/H/9636–15

FIGURE 12. Color Value Write Followed by Any Write

Timing Waveforms (Continued)



TL/H/9636–16

FIGURE 13. Color Value Read Followed by Any Read

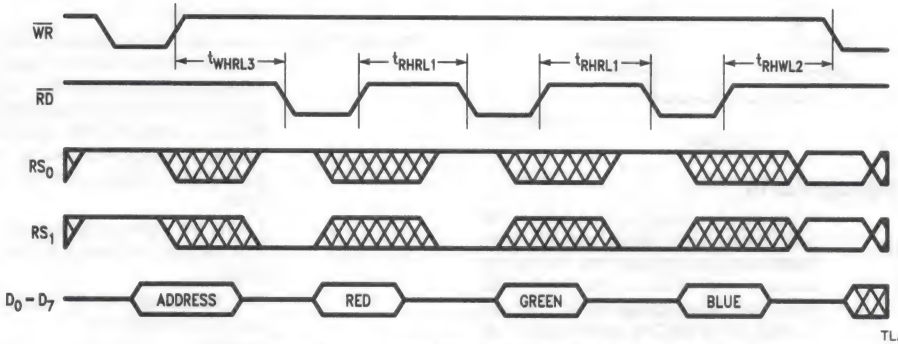


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FIGURE 14. Color Value Read Followed by Any Write

Connection Diagram

Plastic & Cavity Dual-In-Line Package

TL/H/9636–18

Top View

Pin Descriptions

RED (1), GREEN (2), BLUE (3) These are the analog output pins of the 6-bit DACs. The output currents from these pins flow through the terminating resistors and develop the RGB (red, green and blue) voltages that drive the monitor. Each DAC is composed of 63 current sources. The output of each of these current sources is summed together according to the applied 6-bit binary value.

I_{REF} (4) This is the Reference Current input. The current forced out of this pin to ground determines the current sourced by each of the 63 current sources in each of the three 6-bit DACs. Each current source produces 1/30 of I_{REF} when activated by the 6-bit digital input code.

P_0–P_7 (5–12) These are the high-speed Pixel Address inputs. This byte-wide information is latched and masked by the Pixel Mask Register. The resulting value is used as an address of a location in the Color Palette RAM.

PCLK (13) The high-speed Pixel Clock signal is applied to this pin. The rising edge controls the latching of the Pixel Address and Blanking inputs. It also controls the progress of these values through the three stage pipeline of the Color Palette and through the DACs to the outputs.

GND (14) This is the ground power supply connection.

$\overline{RD}$ (15) This is the active low Read bus control signal. When active, any information present on the internal data bus is available on the Data I/O lines, D_0–D_7.

$\overline{BLANK}$ (16) This is an active low signal that forces the DAC's outputs to zero. When $\overline{BLANK}$ is asserted a video monitor's screen becomes black and the DACs ignore any output values from the Color Palette. However, the Color Palette can still be updated through D_0–D_7.

D_0–D_7 (17–24) These are the bidirectional Data I/O lines used by the host microprocessor to write information (using the active low $\overline{WR}$) into and read information (using the active low $\overline{RD}$) from the DAC0630 and DAC0631's internal registers (Pixel Address register, Color Value register, and Pixel Mask register).

During the write cycle, the rising edge of $\overline{WR}$ latches the data into the selected register.

The rising edge of $\overline{RD}$ determines the end of the read cycle.

With $\overline{RD}$ and $\overline{WR}$ equal to a logic high, the Data I/O lines will no longer contain information from the selected register and will go into a tristate mode.

$\overline{WR}$ (25) This is the active low Write signal. It controls the timing of the write operations on the microprocessor interface inputs, D_0–D_7. When active, any information present on the external data bus is available to the Data I/O lines, D_0–D_7.

RS_0, RS_1 (26, 27) These are the Register Select lines which control the selection of one of the three internal registers. These two lines are sampled during the falling edges of the enable signals ($\overline{RD}$ or $\overline{WR}$). See Functional Description for more information regarding the internal registers.

V^+ (28) This is the positive supply pin. It is normally connected to +5 Vdc and bypassed with a 10 μF tantalum capacitor and a 0.1 μF chip capacitor.

Functional Description

The DAC0630 (or DAC0631) forms the output stage for high resolution raster scan RGB video systems. It contains a Color Palette with 256 memory locations that are 18 bits wide. The color palette's output is connected to three high speed current output 6-bit video DACs. The devices use on-board registers to interface easily with microprocessors.

MICROPROCESSOR INTERFACE

The DAC0630 and DAC0631's microprocessor interface consists of three internal registers; Pixel Address register, Color Value register, and Pixel Mask register. These are individually accessed by register select signals, RS_0 and RS_1. The following table defines which of the three internal registers is selected by each of the four combinations of logic states of RS_0 and RS_1.

RS_0	RS_1	Register
0	0	Pixel Address (Write Mode)
1	1	Pixel Address (Read Mode)
1	0	Color Value
0	1	Pixel Mask

The contents of the color palette can be accessed through the Color Value and Pixel Address registers.

All of the operations on the microprocessor interface can take place asynchronously to the pixel information currently being processed by the Color Palette.

The **Pixel Address** register is a byte-wide latch that receives and latches address information applied to pins 17–24. It can be used in either the Read and Write mode depending on the logic state of RS_0 and RS_1. With $RS_0 = RS_1 = 0$ (register select = 0,0), the Pixel address register is in the **write** mode. Two events normally precede *writing* one or more new color definitions to the color palette. The first is the specification of a color palette address. Second, the Color Value register must be loaded with a color definition. The sequence of data transfer is 1) the desired color palette address (this address is stored in the Pixel Address register) and 2) the color definitions: red, green and blue. Refer to *Figures 11* and *12*.

When $RS_0 = RS_1 = 1$ (register select = 1,1), the Pixel Address register is in the **read** mode. Once again, two events take place and normally precede *reading* one or more color definitions in the color palette. The first action is to specify an address within the color palette. the second is to load the Color Value register with the contents of the color palette location whose address is stored in the Pixel Address Register. The color definition data transfer sequence is red, green and blue. Refer to *Figures 10, 13* and *14*.

The **Color Value** register is an internal 18-bit wide register used as a buffer between the microprocessor interface and the color palette. It is accessed by setting $RS_0 = 1$ and $RS_1 = 0$. A color definition can be read from or written to this register by a sequence of three byte-wide transfers to this register address. When a byte is written to this register, only the least significant six bits (D_0–D_5) contain color information. When a byte is read from this register address, only the six least significant bits contain information—the most significant two bits are set to zero. Refer to *Figures 10–14*.

After the write sequence is completed, the Color Value register's contents are written to the specified color palette address stored in the Pixel Address register. Finally, the Pixel Address register is automatically incremented.

It is possible to read the color definitions stored in the DAC's color palette. After setting RS_0 and RS_1 equal to 1, the desired color palette address is stored in the Pixel Address register. The color definition (18-bits) in the desired color palette location is then automatically transferred to the Color Value register and the Pixel Address is auto-incremented. With successive read cycles, the color definitions pointed to by the incremented address are transferred to the color value register. Refer to *Figure 13*.

The **Pixel Mask** register is a byte-wide latch. by setting $RS_0 = 0$ and $RS_1 = 1$, the Pixel Mask register can be accessed by the microprocessor interface, D_0–D_7. This register is used to mask selected bits of the pixel address values applied to the Pixel Address inputs (P_0–P_7). A "1" in any location in the Pixel Mask register leaves the corresponding bit in the pixel address unchanged. A "0" will reset the corresponding bit to zero. The operation of the Pixel Mask register does not affect the address of the color definition when the microprocessor accesses the color palette. The masking operation makes it possible to alter the displayed colors without altering the contents of external video memory or the DAC0630/631's color palette.

WRITING TO THE COLOR PALETTE

A new color definition can be stored in the color palette by first specifying the initial address while in *write* mode ($RS_0 = RS_1 = \overline{WR} = 0$). This address is stored in the Pixel Address register. The initial address is followed by the red, green and blue color definition data ($RS_0 = 1$, $RS_1 = WR = 0$). These three six-bit values are collected together in the Color Value register for a total of 18 bits. The internal logic then transfers this new color definition to the location pointed to by the address stored in the Pixel Address register. As soon as this transfer is completed, the Pixel Address register is auto-incremented. This allows consecutive color palette locations to be updated without the microprocessor specifying each address. All that is necessary is to continue supplying the red, green and blue data for each consecutive address. Refer to *Figures 11* and *12*.

Attempting to update the color palette when $\overline{BLANK}$ is not asserted results in the data from the Color Value register taking precedence over the DAC0630 and DAC0631's bit mapping operation. The output of the three 6-bit DACs will be based on the color definition from the memory location specified by the pixel address register and not the address found on P_0–P_7. This conflict results in the DAC's generating unexpected output levels. This can last as long as two P_{CLK} periods.

READING FROM THE COLOR PALETTE

To read a location in the color palette an address is sent on the Data I/O lines (D_0–D_7) while in read mode ($RS_0 = RS_1 = 1$, $\overline{WR} = 0$) and stored in the Pixel Address register. The color definition in the specified color palette location is then transferred to the Color Value register and the Pixel Address register is auto-incremented. The color definition can now be retrieved with three sequential read operations ($RS_0 = 1$, $RS_1 = \overline{RD} = 0$). The first byte placed on the Data I/O lines contains the red value. The next is green, and the last is blue. The two most significant bits are set to zero in each case. Once again, the Pixel Address register is auto-incremented, and consecutive color palette locations can be read simply by specifying the beginning address and reading the color palette one or more times. Refer to *Figures 10, 13* and *14*.

Functional Description (Continued)

If the Pixel address register is ever updated during a read or write operation, the current data sequence is terminated and a new read or write operation is initialized.

VIDEO PATH

The video path consists of the Pixel Latch and Mask (inputs P_0–P_7), color palette (256 x 18-bit wide RAM), 18-bit wide bus, and an 18-bit wide latch on the inputs of the three 6-bit high-speed video DACs. The video path uses a three clock cycle (P_{CLK}) pipeline for the pixel address and $\overline{BLANK}$ inputs. These signals are latched on the rising edge of P_{CLK}.

At each rising edge of P_{CLK}, the Color Palette address applied to P_0–P_7 is stored in the Pixel Latch and defines a location in the Color Palette. The color definition in that location is then transferred to the three 6-bit DAC input latches.

ANALOG OUTPUTS

The analog outputs are designed to drive 75Ω loads with I_{REF} set to 4.44 mA or 37.5Ω loads with I_{REF} set to 8.88 mA. For both loads the peak-white amplitude is 0.7V.

The analog outputs can be set to zero by using the $\overline{BLANK}$ input. This is an active low signal that forces the analog outputs to ground by placing all zeros on the DACs' inputs. The color definition selected by the pixel address is ignored.

The DAC0630/631's DACs use switched current sources that are summed together, thus generating the output current. Each 6-bit DAC consists of 63 current sources, each of which has a magnitude of $I_{REF}/30$. The digital input code determines the number of current sources that are active and contributing to the total output current. This output current, in conjunction with a termination resistance connected between each DAC output and ground, sets the full-scale magnitude of the output voltage as determined by

$$V_{PEAK\ WHITE} = 2.1(I_{REF})R_L$$

$$V_{BLACK\ LEVEL} = 0V$$

Application Hints

POWER SUPPLY

The DAC0630 and DAC0631 draw large transient currents from the power supply. To ensure proper operation it is necessary to utilize standard high frequency board layout and power supply distribution techniques.

The transient currents drawn by the DAC0630 and DAC0631 dictate that the ac impedance at the supply pins must be kept to a minimum. This is accomplished by using the recommended decoupling capacitors, C1 and C2, as shown in *Figure 15*. These capacitors must have leads that are as short as possible. High frequency decoupling is accomplished with a 0.1 μF chip capacitor, C_1. A bead tantalum, between 10 μF to 47 μF, should be used for C_2.

Differential ground noise can be created when a voltage difference appears between pin 14 and the ground of the digital devices driving the DAC0630 or DAC0631. This voltage difference is caused by series impedance in the ground path and the current transients drawn by the DAC0630 or DAC0631. The differential ground noise can be minimized by using large, low inductance ground paths between the digital devices that drive the DAC0630 or DAC0631 and pin 14. Therefore, a ground plane layout is recommended.

ANALOG OUTPUT—LINE DRIVING

The connection between the DAC's outputs and the RGB inputs of the video monitor it is driving should be viewed as a transmission line. Impedance changes along this line will result in the reflection of part of the video signal back to the DAC's outputs. These reflections may result in a degradation of the picture quality displayed on the monitor.

To ensure good signal fidelity, RF techniques should be observed. Any traces connecting the DAC0630 or DAC0631 to an on-board connector should form a transmission line of 75Ω impedance. However, the need to ensure that the connecting traces form a transmission line can be eliminated by placing the DAC's output termination resistors at the output connector instead of the DAC's output pins.

The coaxial cable that connects the DAC's outputs to a video monitor should have a characteristic impedance of 75Ω. Connectors on the coaxial line can cause impedance change. Any connectors used with the coaxial cable should match its characteristic impedance.

There are four different methods of terminating the DAC outputs:

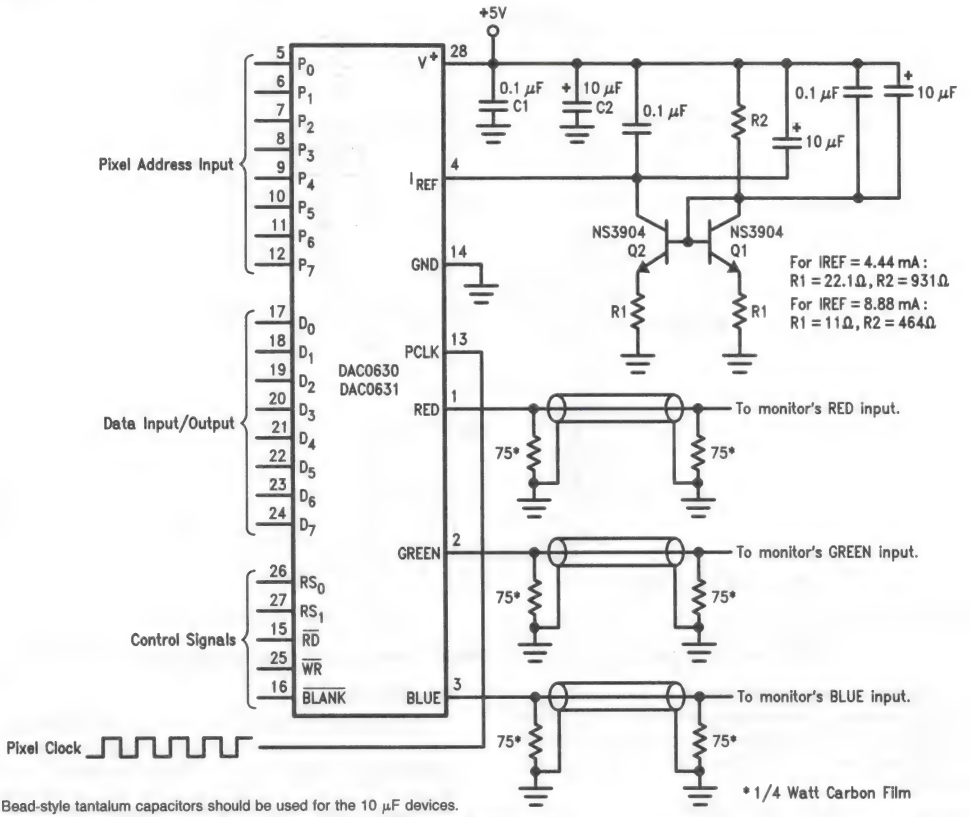
1) Single termination at the DAC (75Ω)
2) Single termination at the destination (75Ω)
3) Double termination (37.5Ω)
4) Buffered signal

1) Single termination at the **source** involves placing a single termination resistor at each DAC output of the DAC0630 and DAC0631 (or at the connector, as described above). No other terminating load is present. Therefore, a high-input impedance monitor should be used. The ac load driven by the DAC's outputs is the transmission line impedance in parallel with the load resistor. The transmission line's impedance should match the impedance of the load resistor. Thus, the DAC's output has an initial signal amplitude that is half the dc value expected. This half-amplitude signal is 100% reflected by the open circuit presented by the monitor input. This restores the signal amplitude to the expected value. The reflections from the monitor propagate back towards the DAC outputs. The load resistor at each DAC output presents a correctly terminated transmission line so no further reflections occur. This arrangement is relatively tolerant to mismatches in the transmission line between the DAC and the monitor because no reflections occur at the DAC end of the transmission line. However, multiple monitors should not be connected in parallel despite each monitor's high input impedance.

2) Single termination at the **destination** has the termination impedance at the input of the monitor acting as both the load resistor for the DAC and the termination impedance of the cable (transmission line). If the connection between the DAC0630/631 is correctly terminated there will be no reflections. However, if there are any line impedance variations along the cable, reflections will occur and create "ghost images" on the display. This occurs because there is a reflection from the point where the mismatch occurs back to the DAC's output. The signal then reflects off the DAC's output back toward the monitor. It arrives with a significant time delay following the original signal, and "ghosting" results.

3) Double termination of the DAC outputs allow each end of the transmission line to be correctly matched. This results in the least amount of reflection and the highest signal and display fidelity. This termination method also allows for the

Application Hints (Continued)



Note: Bead-style tantalum capacitors should be used for the 10 μF devices.
Thermally connect the NPN transistors together with a Wakefield 259 series Equalizing Link.

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FIGURE 15. Typical Connection Showing I_{REF} Generator and Double Termination

fastest fall time. The DAC termination's RC time constant sets the outputs' fall time. The greater the time constant, the slower the fall time. Therefore, the fall time will be minimized since the impedance using this termination technique is less than that achieved with single termination. With double-termination it is necessary to increase I_{REF} to 8.88 mA to ensure a full-scale output voltage of 700 mV.

4) By placing a **buffer** at the DAC's output, the DAC0630 and DAC0631 will be able to drive large capacitive loads such as long lossy cables. The buffer requries a high input impedance, a condition that is satisfied with LM1203 RGB Video Amplifier System. A 75Ω load is placed at the buffer's input. The buffer's low output impedance should be matched to the interconnecting cable with a series resistor. The cable should then be terminated with the same resistance at the monitor.

ANALOG OUTPUT—PROTECTION

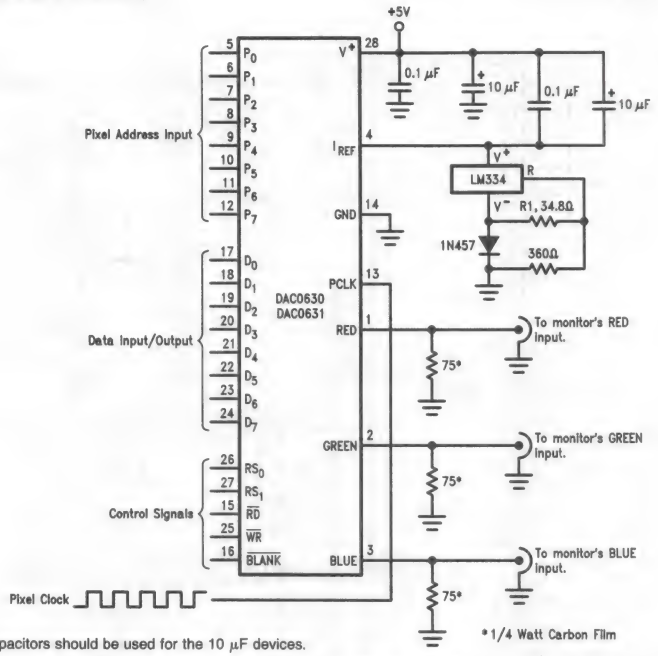
The DAC0630 and DAC0631 have on-chip electrostatic discharge (ESD) protection on each pin. However, the same precautions should be taken as with any other CMOS integrated circuit during manufacturing to reduce the possibility of ESD damage.

GENERATING I_{REF}

An active current source for I_{REF} is recommended to ensure that the DACs have predictable and stable output currents. There are numerous methods available to generate the reference current. The voltage drop from V^+ to the I_{REF} pin increases with increasing I_{REF} current. The circuit used to generate I_{REF} must be designed to operate at the minimum voltage ($V_{REFmin} = V^+ - 3V$) expected from the I_{REF} pin to ground. For any application, V_{REFmin} will be smallest when I_{REF} is maximum and supply voltage is minimum. For $I_{REF} = 8.8$ mA and $V^+ = 4.5V$, the I_{REF} generator will have to operate with 1.5V or less across it. I_{REF} generators that require a voltage drop greater than 1.5V may be used if a negative supply is available.

A simple I_{REF} generator circuit is shown with the DAC0630/DAC0631 in *Figure 15.* As shown, this I_{REF} generator will sink ≈4.44 mA (single termination) with R1 = 22.1Ω and R2 = 931Ω. For applications that use double termination, R1 = 11Ω and R2 = 464Ω. The diode connected transistor, Q1, across Q2's base-emitter junction performs a first-order compensation for thermal variations. It is important to keep the lead lengths as short as possible. This will help reduce stray capacitance and the amount of P_{CLK} that is fed into the I_{REF} pin.

Application Hints (Continued)



Note: Bead-style tantalum capacitors should be used for the 10 μF devices.

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FIGURE 16. Single Termination with LM334 Current Source I_{REF} Generator

Figure 16 shows an alternative method of generating I_{REF}. The LM334 precision current source is used in a temperature compensated configuration. The reference current is set by a single resistor, R1, independent of V^+. The current's value is

$$I_{REF} \approx 160\ mV/R1$$

DECOUPLING I_{REF}

The magnitude of the current flowing through the internal current sources depends not only on I_{REF}, but also on the voltage at pin 4 relative to V^+. Therefore, voltage variations between V^+ and the I_{REF} input can result in variations in the DAC's output current. These variations can be greatly attenuated by using a high frequency capacitor in parallel with a larger electrolytic capacitor to couple the I_{REF} input to V^+.

National Semiconductor

DAC0800/DAC0801/DAC0802 8-Bit Digital-to-Analog Converters

General Description

The DAC0800 series are monolithic 8-bit high-speed current-output digital-to-analog converters (DAC) featuring typical settling times of 100 ns. When used as a multiplying DAC, monotonic performance over a 40 to 1 reference current range is possible. The DAC0800 series also features high compliance complementary current outputs to allow differential output voltages of 20 Vp-p with simple resistor loads as shown in *Figure 1*. The reference-to-full-scale current matching of better than ±1 LSB eliminates the need for full-scale trims in most applications while the nonlinearities of better than ±0.1% over temperature minimizes system error accumulations.

The noise immune inputs of the DAC0800 series will accept TTL levels with the logic threshold pin, V_{LC}, grounded. Changing the V_{LC} potential will allow direct interface to other logic families. The performance and characteristics of the device are essentially unchanged over the full ±4.5V to ±18V power supply range; power dissipation is only 33 mW with ±5V supplies and is independent of the logic input states.

The DAC0800, DAC0802, DAC0800C, DAC0801C and DAC0802C are a direct replacement for the DAC-08, DAC-08A, DAC-08C, DAC-08E and DAC-08H, respectively.

Features

- Fast settling output current — 100 ns
- Full scale error — ±1 LSB
- Nonlinearity over temperature — ±0.1%
- Full scale current drift — ±10 ppm/°C
- High output compliance — −10V to +18V
- Complementary current outputs
- Interface directly with TTL, CMOS, PMOS and others
- 2 quadrant wide range multiplying capability
- Wide power supply range — ±4.5V to ±18V
- Low power consumption — 33 mW at ±5V
- Low cost

Typical Applications

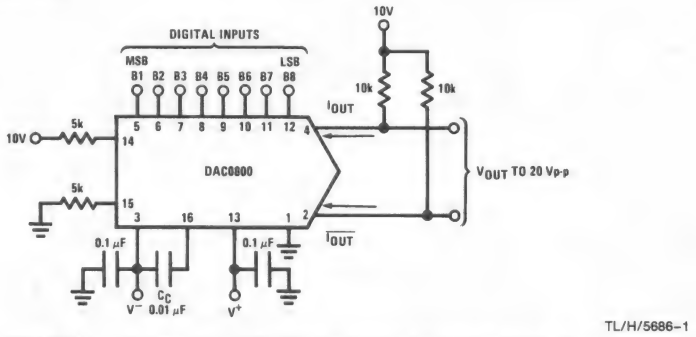

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FIGURE 1. ±20 V_{p-p} Output Digital-to-Analog Converter (Note 4)

Ordering Information

Non-Linearity	Temperature Range	Order Numbers				
		J Package (J16A)*		N Package (N16A)*		SO Package (M16A)
±0.1% FS	−55°C ≤ T_A ≤ +125°C	DAC0802LJ	DAC-08AQ			
±0.1% FS	0°C ≤ T_A ≤ +70°C	DAC0802LCJ	DAC-08HQ	DAC0802LCN	DAC-08HP	DAC0802LCM
±0.19% FS	−55°C ≤ T_A ≤ +125°C	DAC0800LJ	DAC-08Q			
±0.19% FS	0°C ≤ T_A ≤ +70°C	DAC0800LCJ	DAC-08EQ	DAC0800LCN	DAC-08EP	DAC0800LCM
±0.39% FS	0°C ≤ T_A ≤ +70°C	DAC0801LCJ	DAC-08CQ	DAC0801LCN	DAC-08CP	DAC0801LCM

*Devices may be ordered by using either order number.

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage ($V^+ - V^-$)	±18V or 36V
Power Dissipation (Note 2)	500 mW
Reference Input Differential Voltage (V14 to V15)	V^- to V^+
Reference Input Common-Mode Range (V14, V15)	V^- to V^+
Reference Input Current	5 mA
Logic Inputs	V^- to V^- plus 36V
Analog Current Outputs ($V_S^- = -15V$)	4.25 mA
ESD Susceptibility (Note 3)	TBD V
Storage Temperature	−65°C to +150°C
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
Surface Mount Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C

Operating Conditions (Note 1)

	Min	Max	Units
Temperature (T_A)			
DAC0802L	−55	+125	°C
DAC0800L	−55	+125	°C
DAC0800LC	0	+70	°C
DAC0801LC	0	+70	°C
DAC0802LC	0	+70	°C

Electrical Characteristics

The following specifications apply for $V_S = \pm 15V$, $I_{REF} = 2$ mA and $T_{MIN} \le T_A \le T_{MAX}$ unless otherwise specified. Output characteristics refer to both I_{OUT} and $\overline{I_{OUT}}$.

Symbol	Parameter	Conditions	DAC0802L/ DAC0802LC			DAC0800L/ DAC0800LC			DAC0801LC			Units
			Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
	Resolution		8	8	8	8	8	8	8	8	8	Bits
	Monotonicity		8	8	8	8	8	8	8	8	8	Bits
	Nonlinearity				±0.1			±0.19			±0.39	%FS
t_s	Settling Time	To ±½ LSB, All Bits Switched "ON" or "OFF", $T_A = 25°C$		100	135					100	150	ns
		DAC0800L					100	135				ns
		DAC0800LC					100	150				ns
tPLH, tPHL	Propagation Delay	$T_A = 25°C$										
	Each Bit			35	60		35	60		35	60	ns
	All Bits Switched			35	60		35	60		35	60	ns
TCI_{FS}	Full Scale Tempco			±10	±50		±10	±50		±10	±80	ppm/°C
V_{OC}	Output Voltage Compliance	Full Scale Current Change <½ LSB, $R_{OUT} > 20$ MΩ Typ	−10		18	−10		18	−10		18	V
I_{FS4}	Full Scale Current	$V_{REF} = 10.000V$, R14 = 5.000 kΩ R15 = 5.000 kΩ, $T_A = 25°C$	1.984	1.992	2.000	1.94	1.99	2.04	1.94	1.99	2.04	mA
I_{FSS}	Full Scale Symmetry	$I_{FS4} - I_{FS2}$		±0.5	±4.0		±1	±8.0		±2	±16	μA
I_{ZS}	Zero Scale Current			0.1	1.0		0.2	2.0		0.2	4.0	μA
I_{FSR}	Output Current Range	$V^- = -5V$	0	2.0	2.1	0	2.0	2.1	0	2.0	2.1	mA
		$V^- = -8V$ to $-18V$	0	2.0	4.2	0	2.0	4.2	0	2.0	4.2	mA
	Logic Input Levels											
V_{IL}	Logic "0"	$V_{LC} = 0V$			0.8			0.8			0.8	V
V_{IH}	Logic "1"		2.0			2.0			2.0			V
	Logic Input Current	$V_{LC} = 0V$										
I_{IL}	Logic "0"	$-10V \le V_{IN} \le +0.8V$		−2.0	−10		−2.0	−10		−2.0	−10	μA
I_{IH}	Logic "1"	$2V \le V_{IN} \le +18V$		0.002	10		0.002	10		0.002	10	μA
V_{IS}	Logic Input Swing	$V^- = -15V$	−10		18	−10		18	−10		18	V
V_{THR}	Logic Threshold Range	$V_S = \pm 15V$	−10		13.5	−10		13.5	−10		13.5	V
I_{15}	Reference Bias Current			−1.0	−3.0		−1.0	−3.0		−1.0	−3.0	μA
dI/dt	Reference Input Slew Rate	*(Figure 12)*	4.0	8.0		4.0	8.0		4.0	8.0		mA/μs
$PSSI_{FS+}$	Power Supply Sensitivity	$4.5V \le V+ \le 18V$		0.0001	0.01		0.0001	0.01		0.0001	0.01	%/%
$PSSI_{FS-}$		$-4.5V \le V^- \le 18V$ $I_{REF} = 1$mA		0.0001	0.01		0.0001	0.01		0.0001	0.01	%/%
	Power Supply Current	$V_S = \pm 5V$, $I_{REF} = 1$ mA										
I+				2.3	3.8		2.3	3.8		2.3	3.8	mA
I−				−4.3	−5.8		−4.3	−5.8		−4.3	−5.8	mA
		$V_S = 5V$, $-15V$, $I_{REF} = 2$ mA										
I+				2.4	3.8		2.4	3.8		2.4	3.8	mA
I−				−6.4	−7.8		−6.4	−7.8		−6.4	−7.8	mA
		$V_S = \pm 15V$, $I_{REF} = 2$ mA										
I+				2.5	3.8		2.5	3.8		2.5	3.8	mA
I−				−6.5	−7.8		−6.5	−7.8		−6.5	−7.8	mA

Electrical Characteristics (Continued)

The following specifications apply for $V_S = \pm 15V$, $I_{REF} = 2$ mA and $T_{MIN} \leq T_A \leq T_{MAX}$ unless otherwise specified. Output characteristics refer to both I_{OUT} and $\overline{I_{OUT}}$.

Symbol	Parameter	Conditions	DAC0802L/ DAC0802LC			DAC0800L/ DAC0800LC			DAC0801LC			Units
			Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
P_D	Power Dissipation	$\pm 5V$, $I_{REF} = 1$ mA		33	48		33	48		33	48	mW
		$5V, -15V$, $I_{REF} = 2$ mA		108	136		108	136		108	136	mW
		$\pm 15V$, $I_{REF} = 2$ mA		135	174		135	174		135	174	mW

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

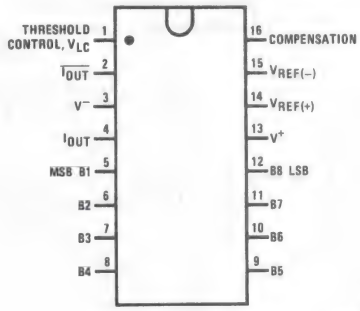
Note 2: The maximum junction temperature of the DAC0800, DAC0801 and DAC0802 is 125°C. For operating at elevated temperatures, devices in the Dual-In-Line J package must be derated based on a thermal resistance of 100°C/W, junction-to-ambient, 175°C/W for the molded Dual-In-Line N package and 100°C/W for the Small Outline M package.

Note 3: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 4: Pin-out numbers for the DAC080X represent the Dual-In-Line package. The Small Outline package pin-out differs from the Dual-In-Line package.

Connection Diagrams

Dual-In-Line Package



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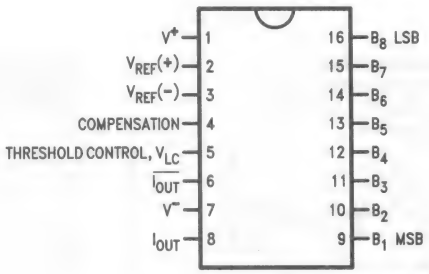
Top View

Small Outline Package



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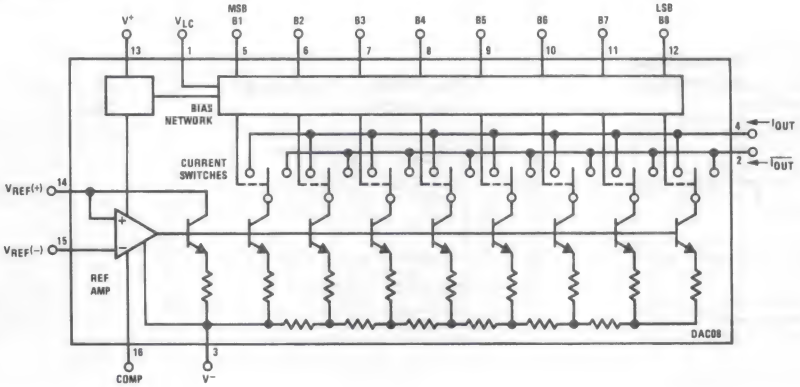
Top View

See Ordering Information

Block Diagram (Note 4)



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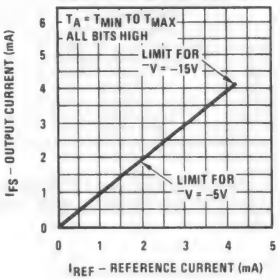
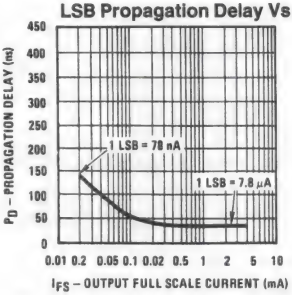
Typical Performance Characteristics

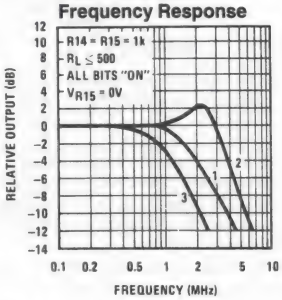
Full Scale Current vs Reference Current



LSB Propagation Delay Vs I_{FS}



Reference Input Frequency Response

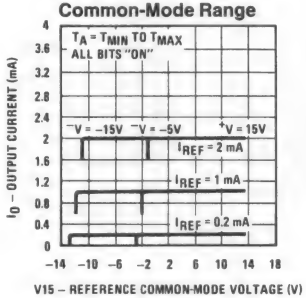


Curve 1: $C_C = 15$ pF, $V_{IN} = 2$ Vp-p centered at 1V.

Curve 2: $C_C = 15$ pF, $V_{IN} = 50$ mVp-p centered at 200 mV.

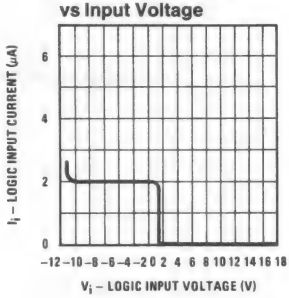
Curve 3: $C_C = 0$ pF, $V_{IN} = 100$ mVp-p at 0V and applied through 50 Ω connected to pin 14.2V applied to R14.

Reference Amp Common-Mode Range

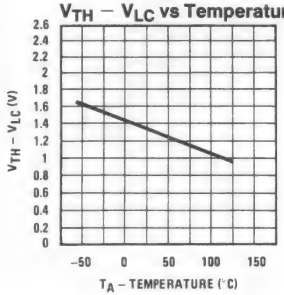

$T_A = T_{MIN}$ TO T_{MAX}
ALL BITS "ON"
−V = −15V −V = −5V +V = 15V
I_{REF} = 2 mA
I_{REF} = 1 mA
I_{REF} = 0.2 mA
I_O – OUTPUT CURRENT (mA)
V15 – REFERENCE COMMON-MODE VOLTAGE (V)

Note. Positive common-mode range is always (V+) − 1.5V

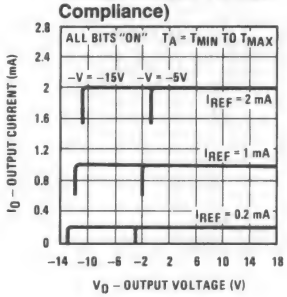
Logic Input Current vs Input Voltage



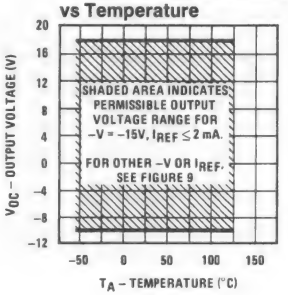
$V_{TH} - V_{LC}$ vs Temperature



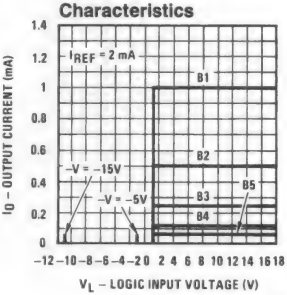
Output Current vs Output Voltage (Output Voltage Compliance)



Output Voltage Compliance vs Temperature


SHADED AREA INDICATES PERMISSIBLE OUTPUT VOLTAGE RANGE FOR −V = −15V, $I_{REF} \leq 2$ mA.
FOR OTHER −V OR I_{REF}, SEE FIGURE 9
V_{OC} – OUTPUT VOLTAGE (V)
T_A – TEMPERATURE (°C)

Bit Transfer Characteristics



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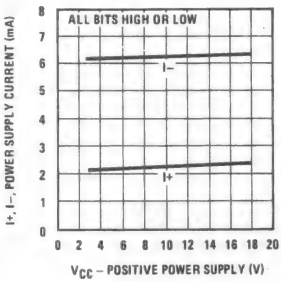
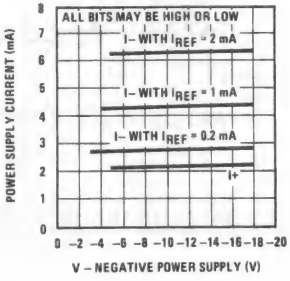
Note. B1–B8 have identical transfer characteristics. Bits are fully switched with less than ½ LSB error, at less than ±100 mV from actual threshold. These switching points are guaranteed to lie between 0.8 and 2V over the operating temperature range (V_{LC} = 0V).

Typical Performance Characteristics (Continued)

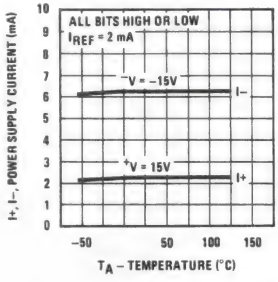
Power Supply Current vs $+V$



Power Supply Current vs $-V$

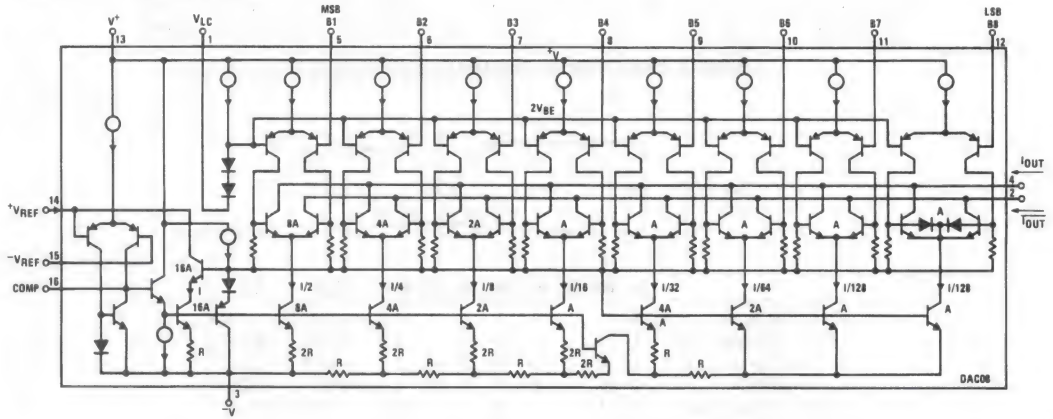

ALL BITS MAY BE HIGH OR LOW
I− WITH IREF = 2 mA
I− WITH IREF = 1 mA
I− WITH IREF = 0.2 mA
I+
POWER SUPPLY CURRENT (mA)
V − NEGATIVE POWER SUPPLY (V)

Power Supply Current vs Temperature



TL/H/5686–4

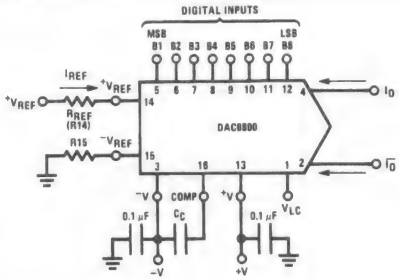
Equivalent Circuit



TL/H/5686–15

FIGURE 2

Typical Applications (Continued)



TL/H/5686–5

$$I_{FS} \approx \frac{+V_{REF}}{R_{REF}} \times \frac{255}{256}$$

$I_O + \overline{I_O} = I_{FS}$ for all logic states

For fixed reference, TTL operation, typical values are:

$V_{REF} = 10.000V$

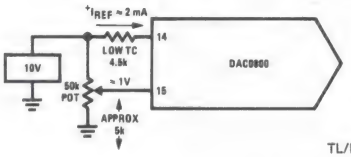
$R_{REF} = 5.000k$

$R15 \approx R_{REF}$

$C_C = 0.01\ \mu F$

$V_{LC} = 0V$ (Ground)

FIGURE 3. Basic Positive Reference Operation (Note 4)



TL/H/5686–21

FIGURE 4. Recommended Full Scale Adjustment Circuit (Note 4)

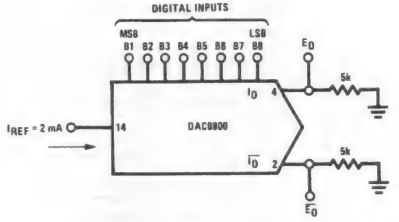
TL/H/5686–16

$$I_{FS} \approx \frac{-V_{REF}}{R_{REF}} \times \frac{255}{256}$$

Note. R_{REF} sets I_{FS}; R15 is for bias current cancellation

FIGURE 5. Basic Negative Reference Operation (Note 4)

Typical Applications (Continued)



TL/H/5686–17

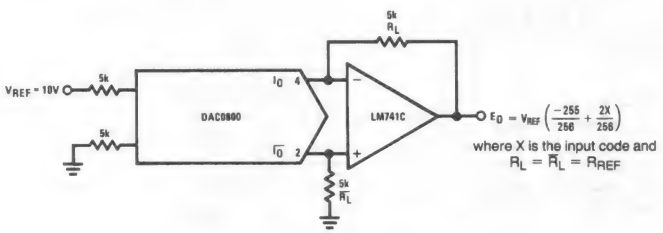
	B1	B2	B3	B4	B5	B6	B7	B8	I_O mA	$\overline{I_O}$ mA	E_O	$\overline{E_O}$
Full Scale	1	1	1	1	1	1	1	1	1.992	0.000	−9.960	0.000
Full Scale − LSB	1	1	1	1	1	1	1	0	1.984	0.008	−9.920	−0.040
Half Scale + LSB	1	0	0	0	0	0	0	1	1.008	0.984	−5.040	−4.920
Half Scale	1	0	0	0	0	0	0	0	1.000	0.992	−5.000	−4.960
Half Scale − LSB	0	1	1	1	1	1	1	1	0.992	1.000	−4.960	−5.000
Zero Scale + LSB	0	0	0	0	0	0	0	1	0.008	1.984	−0.040	−9.920
Zero Scale	0	0	0	0	0	0	0	0	0.000	1.992	0.000	−9.960

FIGURE 6. Basic Unipolar Negative Operation (Note 4)

TL/H/5686–6

	B1	B2	B3	B4	B5	B6	B7	B8	E_O	$\overline{E_O}$
Pos. Full Scale	1	1	1	1	1	1	1	1	−9.920	+10.000
Pos. Full Scale − LSB	1	1	1	1	1	1	1	0	−9.840	+9.920
Zero Scale + LSB	1	0	0	0	0	0	0	1	−0.080	+0.160
Zero Scale	1	0	0	0	0	0	0	0	0.000	+0.080
Zero Scale − LSB	0	1	1	1	1	1	1	1	+0.080	0.000
Neg. Full Scale + LSB	0	0	0	0	0	0	0	1	+9.920	−9.840
Neg. Full Scale	0	0	0	0	0	0	0	0	+10.000	−9.920

FIGURE 7. Basic Bipolar Output Operation (Note 4)



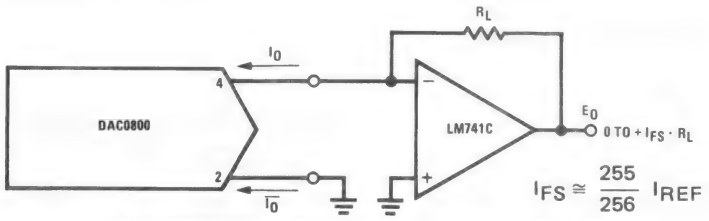
TL/H/5686–18

If $R_L = \overline{R_L}$ within ±0.05%, output is symmetrical about ground

	B1	B2	B3	B4	B5	B6	B7	B8	E_O
Pos. Full Scale	1	1	1	1	1	1	1	1	+9.960
Pos. Full Scale − LSB	1	1	1	1	1	1	1	0	+9.880
(+)Zero Scale	1	0	0	0	0	0	0	0	+0.040
(−)Zero Scale	0	1	1	1	1	1	1	1	−0.040
Neg. Full Scale + LSB	0	0	0	0	0	0	0	1	−9.880
Neg. Full Scale	0	0	0	0	0	0	0	0	−9.960

FIGURE 8. Symmetrical Offset Binary Operation (Note 4)

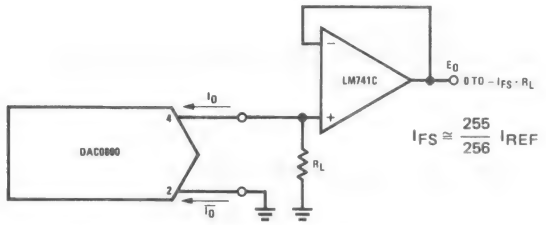
Typical Applications (Continued)



TL/H/5686–19

For complementary output (operation as negative logic DAC), connect inverting input of op amp to $\overline{I_O}$ (pin 2), connect I_O (pin 4) to ground.

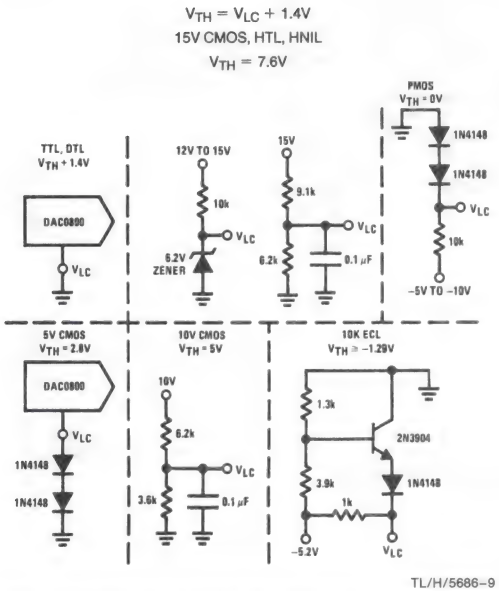
FIGURE 9. Positive Low Impedance Output Operation (Note 4)



TL/H/5686–20

For complementary output (operation as a negative logic DAC) connect non-inverting input of op am to $\overline{I_O}$ (pin 2); connect I_O (pin 4) to ground.

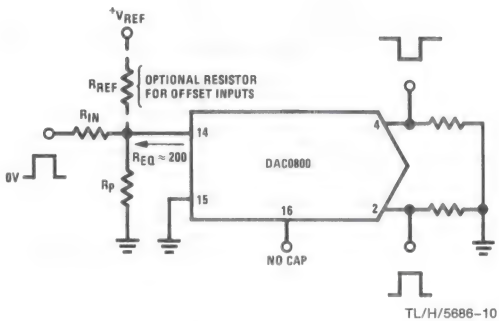
FIGURE 10. Negative Low Impedance Output Operation (Note 4)



TL/H/5686–9

Note. Do not exceed negative logic input range of DAC.

FIGURE 11. Interfacing with Various Logic Families

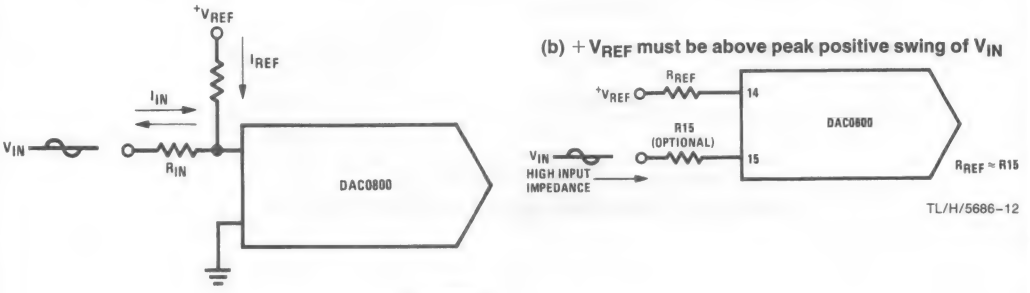


TL/H/5686–10

Typical values: $R_{IN} = 5k, +V_{IN} = 10V$

FIGURE 12. Pulsed Reference Operation (Note 4)

Typical Applications (Continued)



TL/H/5686–11

FIGURE 13. Accommodating Bipolar References (Note 4)

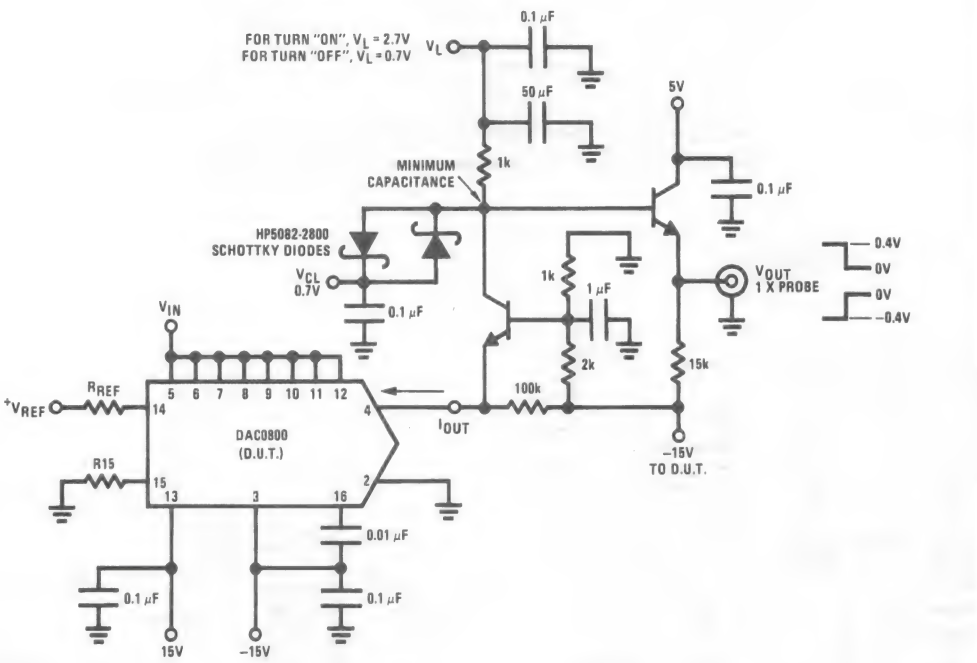
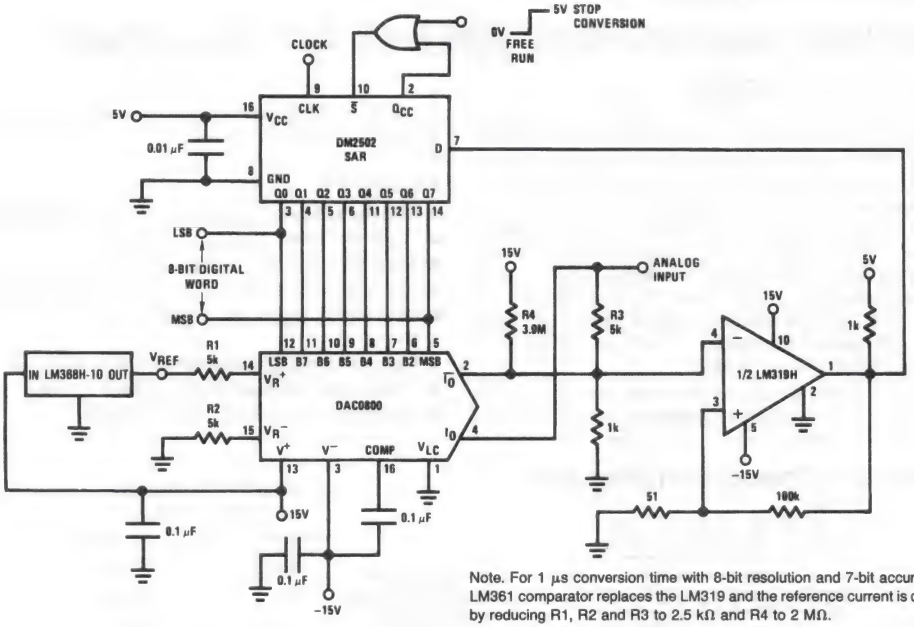


FIGURE 14. Settling Time Measurement (Note 4)

Typical Applications (Continued)



Note. For 1 μs conversion time with 8-bit resolution and 7-bit accuracy, an LM361 comparator replaces the LM319 and the reference current is doubled by reducing R1, R2 and R3 to 2.5 kΩ and R4 to 2 MΩ.

TL/H/5686–8

FIGURE 15. A Complete 2 μs Conversion Time, 8-Bit A/D Converter (Note 4)

National Semiconductor

DAC0808/DAC0807/DAC0806 8-Bit D/A Converters

General Description

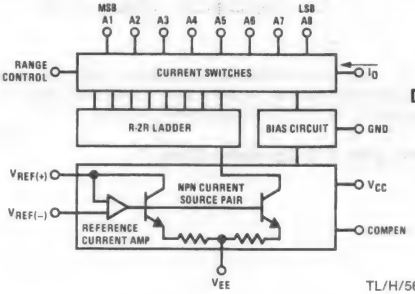
The DAC0808 series is an 8-bit monolithic digital-to-analog converter (DAC) featuring a full scale output current settling time of 150 ns while dissipating only 33 mW with ±5V supplies. No reference current (I_{REF}) trimming is required for most applications since the full scale output current is typically ±1 LSB of 255 I_{REF}/ 256. Relative accuracies of better than ±0.19% assure 8-bit monotonicity and linearity while zero level output current of less than 4 μA provides 8-bit zero accuracy for $I_{REF} \geq 2$ mA. The power supply currents of the DAC0808 series are independent of bit codes, and exhibits essentially constant device characteristics over the entire supply voltage range.

The DAC0808 will interface directly with popular TTL, DTL or CMOS logic levels, and is a direct replacement for the MC1508/MC1408. For higher speed applications, see DAC0800 data sheet.

Features

- Relative accuracy: ±0.19% error maximum (DAC0808)
- Full scale current match: ±1 LSB typ
- 7 and 6-bit accuracy available (DAC0807, DAC0806)
- Fast settling time: 150 ns typ
- Noninverting digital inputs are TTL and CMOS compatible
- High speed multiplying input slew rate: 8 mA/μs
- Power supply voltage range: ±4.5V to ±18V
- Low power consumption: 33 mW @ ±5V

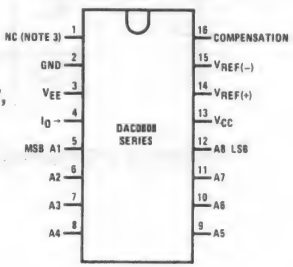
Block and Connection Diagrams



TL/H/5687–1

Order Number DAC0808, DAC0807, or DAC0806 See NS Package Number J16A, M16A or N16A

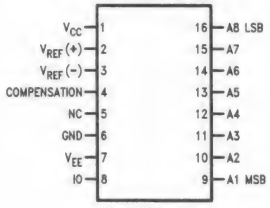
Dual-In-Line Package



TOP VIEW

TL/H/5687–2

Small-Outline Package



TL/H/5687–13

Top View

Ordering Information

ACCURACY	OPERATING TEMPERATURE RANGE	ORDER NUMBERS				
		J PACKAGE (J16A)*		N PACKAGE (N16A)*		SO PACKAGE (M16A)
8-bit	$-55°C \leq T_A \leq +125°C$	DAC0808LJ	MC1508L8			
8-bit	$0°C \leq T_A \leq +75°C$	DAC0808LCJ	MC1408L8	DAC0808LCN	MC1408P8	DAC0808LCM
7-bit	$0°C \leq T_A \leq +75°C$	DAC0807LCJ	MC1408L7	DAC0807LCN	MC1408P7	DAC0807LCM
6-bit	$0°C \leq T_A \leq +75°C$	DAC0806LCJ	MC1408L6	DAC0806LCN	MC1408P6	DAC0806LCM

***Note.** Devices may be ordered by using either order number.

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Power Supply Voltage	
V_{CC}	$+18\ V_{DC}$
V_{EE}	$-18\ V_{DC}$
Digital Input Voltage, V5–V12	$-10\ V_{DC}$ to $+18\ V_{DC}$
Applied Output Voltage, V_O	$-11\ V_{DC}$ to $+18\ V_{DC}$
Reference Current, I_{14}	5 mA
Reference Amplifier Inputs, V14, V15	V_{CC}, V_{EE}
Power Dissipation (Note 3)	1000 mW
ESD Susceptibility (Note 4)	TBD
Storage Temperature Range	−65°C to +150°C
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Surface Mount Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C

Operating Ratings

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
DAC0808L	$-55°C \le T_A \le +125°C$
DAC0808LC Series	$0 \le T_A \le +75°C$

Electrical Characteristics

(V_{CC} = 5V, V_{EE} = −15 V_{DC}, $V_{REF}/R14$ = 2 mA, DAC0808: T_A = −55°C to +125°C, DAC0808C, DAC0807C, DAC0806C, T_A = 0°C to +75°C, and all digital inputs at high logic level unless otherwise noted.)

Symbol	Parameter	Conditions	Min	Typ	Max	Units
E_r	Relative Accuracy (Error Relative to Full Scale I_O)	*(Figure 4)*				%
	DAC0808L (LM1508-8), DAC0808LC (LM1408-8)				±0.19	%
	DAC0807LC (LM1408-7), (Note 5)				±0.39	%
	DAC0806LC (LM1408-6), (Note 5)				±0.78	%
	Settling Time to Within ½ LSB (Includes t_{PLH})	T_A = 25°C (Note 6), *(Figure 5)*		150		ns
t_{PLH}, t_{PHL}	Propagation Delay Time	T_A = 25°C, *(Figure 5)*		30	100	ns
TCI_O	Output Full Scale Current Drift			±20		ppm/°C
MSB	Digital Input Logic Levels	*(Figure 3)*				
V_{IH}	High Level, Logic "1"		2			V_{DC}
V_{IL}	Low Level, Logic "0"				0.8	V_{DC}
MSB	Digital Input Current	*(Figure 3)*				
	High Level	V_{IH} = 5V		0	0.040	mA
	Low Level	V_{IL} = 0.8V		−0.003	−0.8	mA
I_{15}	Reference Input Bias Current	*(Figure 3)*		−1	−3	μA
	Output Current Range	*(Figure 3)*				
		V_{EE} = −5V	0	2.0	2.1	mA
		V_{EE} = −15V, T_A = 25°C	0	2.0	4.2	mA
I_O	Output Current	V_{REF} = 2.000V, R14 = 1000Ω, *(Figure 3)*	1.9	1.99	2.1	mA
	Output Current, All Bits Low	*(Figure 3)*		0	4	μA
	Output Voltage Compliance (Note 2)	$E_r \le 0.19\%$, T_A = 25°C				
	V_{EE} = −5V, I_{REF} = 1 mA				−0.55, +0.4	V_{DC}
	V_{EE} Below −10V				−5.0, +0.4	V_{DC}

Electrical Characteristics (Continued)

(V_{CC} = 5V, V_{EE} = −15 V_{DC}, $V_{REF}/R14$ = 2 mA, DAC0808: T_A = −55°C to +125°C, DAC0808C, DAC0807C, DAC0806C, T_A = 0°C to +75°C, and all digital inputs at high logic level unless otherwise noted.)

Symbol	Parameter	Conditions	Min	Typ	Max	Units
SRI_{REF}	Reference Current Slew Rate	*(Figure 6)*	4	8		mA/μs
	Output Current Power Supply Sensitivity	$-5V \le V_{EE} \le -16.5V$		0.05	2.7	μA/V
	Power Supply Current (All Bits Low)	*(Figure 3)*				
I_{CC}				2.3	22	mA
I_{EE}				−4.3	−13	mA
	Power Supply Voltage Range	T_A = 25°C, *(Figure 3)*				
V_{CC}			4.5	5.0	5.5	V_{DC}
V_{EE}			−4.5	−15	−16.5	V_{DC}
	Power Dissipation					
	All Bits Low	V_{CC} = 5V, V_{EE} = −5V		33	170	mW
		V_{CC} = 5V, V_{EE} = −15V		106	305	mW
	All Bits High	V_{CC} = 15V, V_{EE} = −5V		90		mW
		V_{CC} = 15V, V_{EE} = −15V		160		mW

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: Range control is not required.

Note 3: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\theta_{JA}$ or the number given in the Absolute Maixmum Ratings, whichever is lower. For this device, T_{JMAX} = 125°C, and the typical junction-to-ambient thermal resistance of the dual-in-line J package when the board mounted is 100°C/W. For the dual-in-line N package, this number increases to 175°C/W and for the small outline M package this number is 100°C/W.

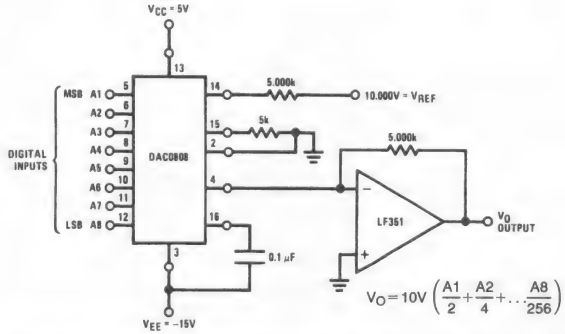
Note 4: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 5: All current switches are tested to guarantee at least 50% of rated current.

Note 6: All bits switched.

Note 7: Pin-out numbers for the DAL080X represent the dual-in-line package. The small outline package pinout differs from the dual-in-line package.

Typical Application



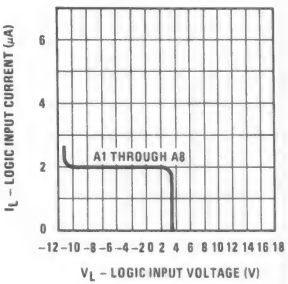
TL/H/5687–3

FIGURE 1. + 10V Output Digital to Analog Converter (Note 7)

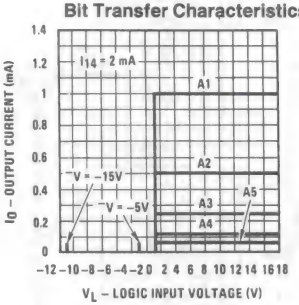
Typical Performance Characteristics

V_{CC} = 5V, V_{EE} = −15V, T_A = 25°C, unless otherwise noted

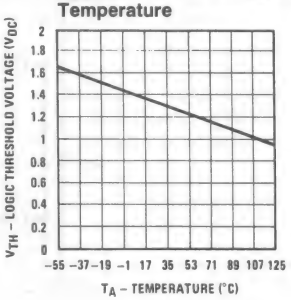
Logic Input Current vs Input Voltage



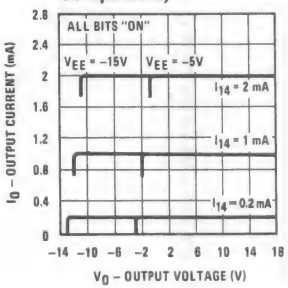
Bit Transfer Characteristics



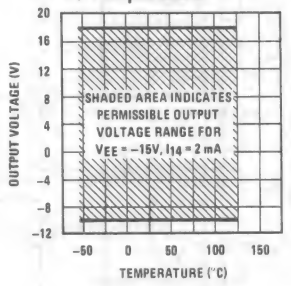
Logic Threshold Voltage vs Temperature



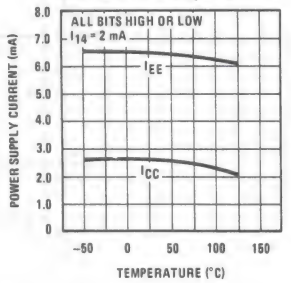
Output Current vs Output Voltage (Output Voltage Compliance)



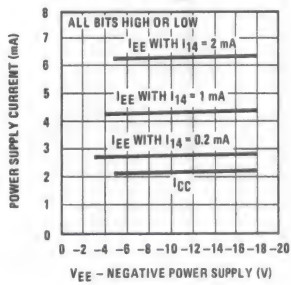
Output Voltage Compliance vs Temperature



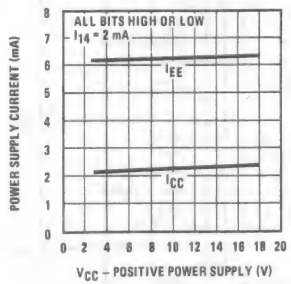
Typical Power Supply Current vs Temperature



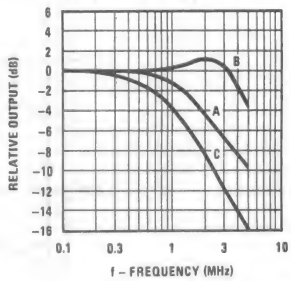
Typical Power Supply Current vs V_{EE}


ALL BITS HIGH OR LOW
I_EE WITH I14 = 2 mA
I_EE WITH I14 = 1 mA
I_EE WITH I14 = 0.2 mA
I_CC
POWER SUPPLY CURRENT (mA)
V_EE – NEGATIVE POWER SUPPLY (V)

Typical Power Supply Current vs V_{CC}



Reference Input Frequency Response



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Unless otherwise specified: R14 = R15 = 1 kΩ, C = 15 pF, pin 16 to V_{EE}; R_L = 50Ω, pin 4 to ground.

Curve A: Large Signal Bandwidth Method of *Figure 7*, V_{REF} = 2 Vp-p offset 1 V above ground.

Curve B: Small Signal Bandwidth Method of *Figure 7*, R_L = 250Ω, V_{REF} = 50 mVp-p offset 200 mV above ground.

Curve C: Large and Small Signal Bandwidth Method of *Figure 9* (no op amp, R_L = 50Ω), R_S = 50Ω, V_{REF} = 2V, V_S = 100 mVp-p centered at 0V.

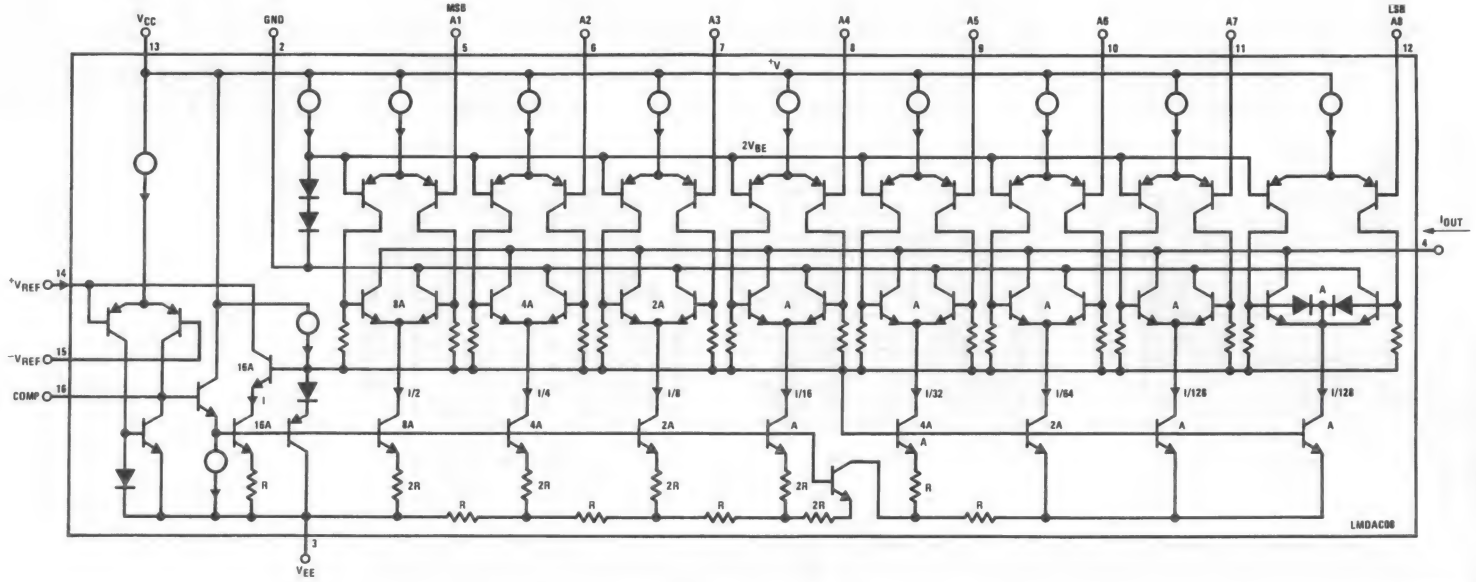

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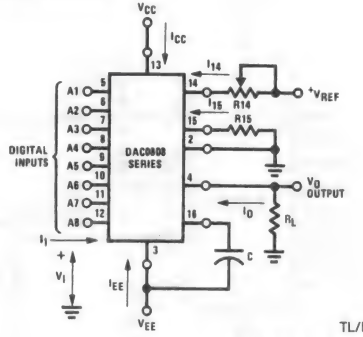
FIGURE 2. Equivalent Circuit of the DAC0808 Series (Note 7)

Test Circuits



V_I and I_1 apply to inputs A1–A8.

The resistor tied to pin 15 is to temperature compensate the bias current and may not be necessary for all applications.

$$I_O = K\left(\frac{A1}{2} + \frac{A2}{4} + \frac{A4}{16} + \frac{A5}{32} + \frac{A6}{64} + \frac{A7}{128} + \frac{A8}{256}\right)$$

where $K \cong \frac{V_{REF}}{R14}$

and A_N = "1" if A_N is at high level

A_N = "0" if A_N is at low level

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FIGURE 3. Notation Definitions Test Circuit (Note 7)

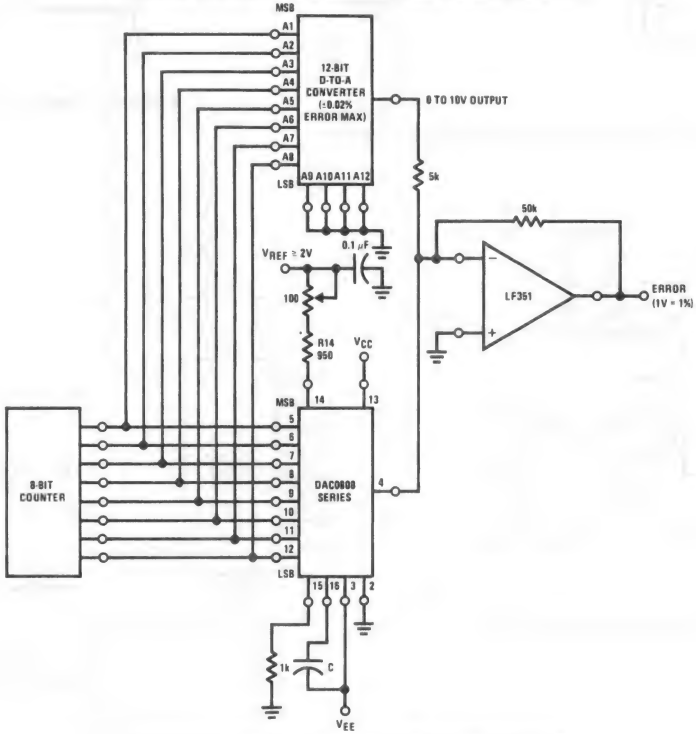

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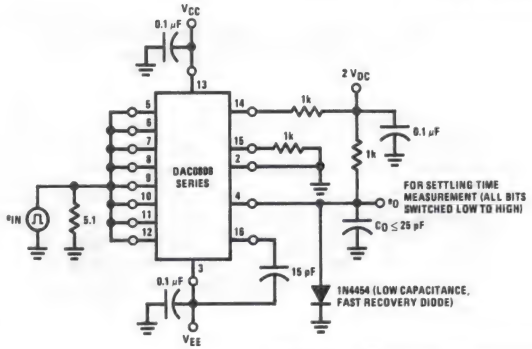
FIGURE 4. Relative Accuracy Test Circuit (Note 7)



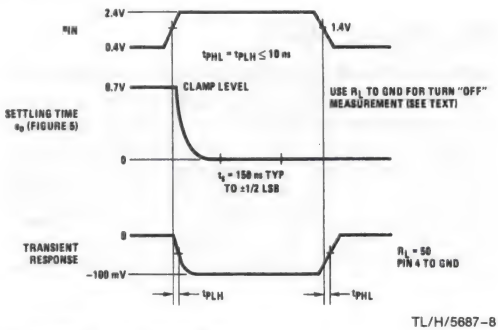


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FIGURE 5. Transient Response and Settling Time (Note 7)

Test Circuits (Continued)

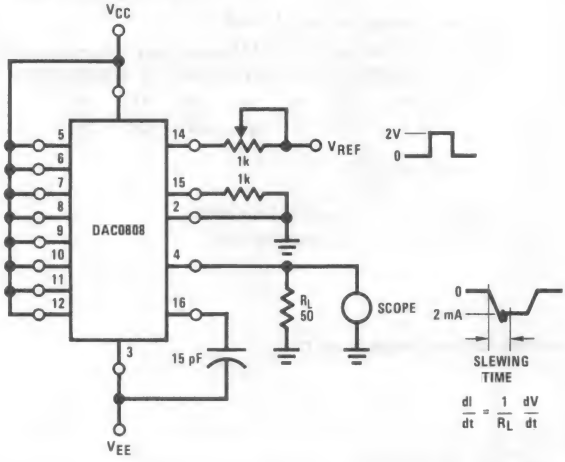

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FIGURE 6. Reference Current Slew Rate Measurement (Note 7)

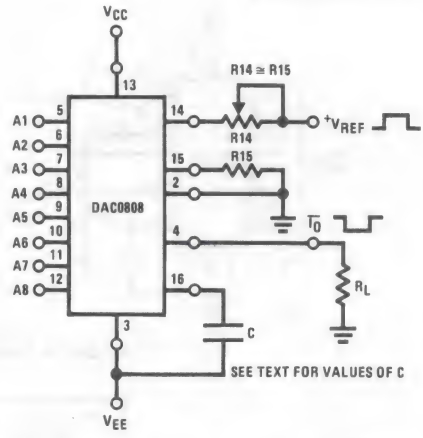

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FIGURE 7. Positive V_{REF} (Note 7)

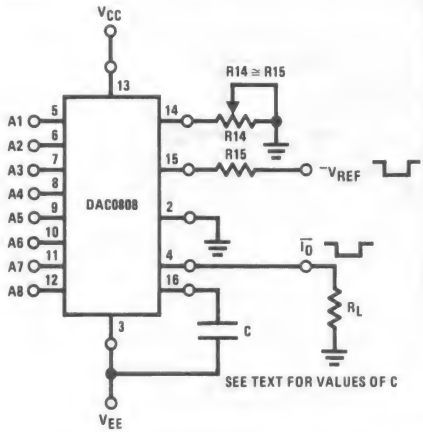

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FIGURE 8. Negative V_{REF} (Note 7)

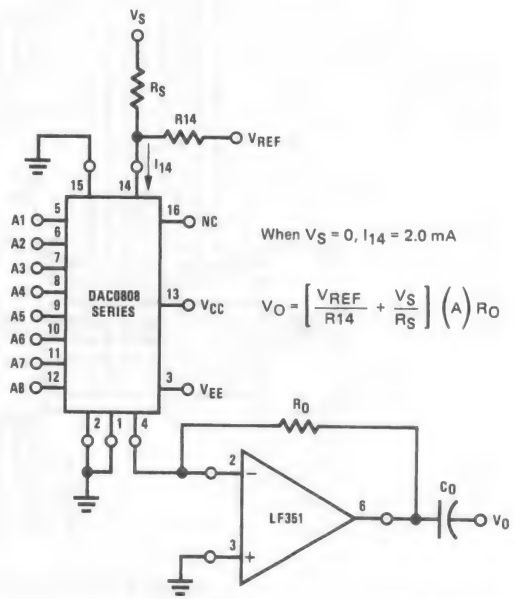

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FIGURE 9. Programmable Gain Amplifier or Digital Attenuator Circuit (Note 7)

Application Hints

REFERENCE AMPLIFIER DRIVE AND COMPENSATION

The reference amplifier provides a voltage at pin 14 for converting the reference voltage to a current, and a turn-around circuit or current mirror for feeding the ladder. The reference amplifier input currrent, I_{14}, must always flow into pin 14, regardless of the set-up method or reference voltage polarity.

Connections for a positive voltage are shown in *Figure 7*. The reference voltage source supplies the full current I_{14}. For bipolar reference signals, as in the multiplying mode, R15 can be tied to a negative voltage corresponding to the minimum input level. It is possible to eliminate R15 with only a small sacrifice in accuracy and temperature drift.

The compensation capacitor value must be increased with increases in R14 to maintain proper phase margin; for R14 values of 1, 2.5 and 5 kΩ, minimum capacitor values are 15, 37 and 75 pF. The capacitor may be tied to either V_{EE} or ground, but using V_{EE} increases negative supply rejection.

Application Hints (Continued)

A negative reference voltage may be used if R14 is grounded and the reference voltage is applied to R15 as shown in *Figure 8*. A high input impedance is the main advantage of this method. Compensation involves a capacitor to V_{EE} on pin 16, using the values of the previous paragraph. The negative reference voltage must be at least 4V above the V_{EE} supply. Bipolar input signals may be handled by connecting R14 to a positive reference voltage equal to the peak positive input level at pin 15.

When a DC reference voltage is used, capacitive bypass to ground is recommended. The 5V logic supply is not recommended as a reference voltage. If a well regulated 5V supply which drives logic is to be used as the reference, R14 should be decoupled by connecting it to 5V through another resistor and bypassing the junction of the 2 resistors with 0.1 μF to ground. For reference voltages greater than 5V, a clamp diode is recommended between pin 14 and ground.

If pin 14 is driven by a high impedance such as a transistor current source, none of the above compensation methods apply and the amplifier must be heavily compensated, decreasing the overall bandwidth.

OUTPUT VOLTAGE RANGE

The voltage on pin 4 is restricted to a range of −0.55 to 0.4V when $V_{EE} = -5V$ due to the current switching methods employed in the DAC0808.

The negative output voltage compliance of the DAC0808 is extended to −5V where the negative supply voltage is more negative than −10V. Using a full-scale current of 1.992 mA and load resistor of 2.5 kΩ between pin 4 and ground will yield a voltage output of 256 levels between 0 and −4.980V. Floating pin 1 does not affect the converter speed or power dissipation. However, the value of the load resistor determines the switching time due to increased voltage swing. Values of R_L up to 500Ω do not significantly affect performance, but a 2.5 kΩ load increases worst-case settling time to 1.2 μs (when all bits are switched ON). Refer to the subsequent text section on Settling Time for more details on output loading.

OUTPUT CURRENT RANGE

The output current maximum rating of 4.2 mA may be used only for negative supply voltages more negative than −8V, due to the increased voltage drop across the resistors in the reference current amplifier.

ACCURACY

Absolute accuracy is the measure of each output current level with respect to its intended value, and is dependent upon relative accuracy and full-scale current drift. Relative accuracy is the measure of each output current level as a fraction of the full-scale current. The relative accuracy of the DAC0808 is essentially constant with temperature due to the excellent temperature tracking of the monolithic resistor ladder. The reference current may drift with temperature, causing a change in the absolute accuracy of output current. However, the DAC0808 has a very low full-scale current drift with temperature.

The DAC0808 series is guaranteed accurate to within ±½ LSB at a full-scale output current of 1.992 mA. This corresponds to a reference amplifier output current drive to the ladder network of 2 mA, with the loss of 1 LSB (8 μA) which is the ladder remainder shunted to ground. The input current to pin 14 has a guaranteed value of between 1.9 and 2.1 mA, allowing some mismatch in the NPN current source pair. The accuracy test circuit is shown in *Figure 4*. The 12-bit converter is calibrated for a full-scale output current of 1.992 mA. This is an optional step since the DAC0808 accuracy is essentially the same between 1.5 and 2.5 mA. Then the DAC0808 circuits' full-scale current is trimmed to the same value with R14 so that a zero value appears at the error amplifier output. The counter is activated and the error band may be displayed on an oscilloscope, detected by comparators, or stored in a peak detector.

Two 8-bit D-to-A converters may not be used to construct a 16-bit accuracy D-to-A converter. 16-bit accuracy implies a total error of ±½ of one part in 65,536 or ±0.00076%, which is much more accurate than the ±0.019% specification provided by the DAC0808.

MULTIPLYING ACCURACY

The DAC0808 may be used in the multiplying mode with 8-bit accuracy when the reference current is varied over a range of 256:1. If the reference current in the multiplying mode ranges from 16 μA to 4 mA, the additional error contributions are less than 1.6 μA. This is well within 8-bit accuracy when referred to full-scale.

A monotonic converter is one which supplies an increase in current for each increment in the binary word. Typically, the DAC0808 is monotonic for all values of reference current above 0.5 mA. The recommended range for operation with a DC reference current is 0.5 to 4 mA.

SETTLING TIME

The worst-case switching condition occurs when all bits are switched ON, which corresponds to a low-to-high transition for all bits. This time is typically 150 ns for settling to within ±½ LSB, for 8-bit accuracy, and 100 ns to ½ LSB for 7 and 6-bit accuracy. The turn OFF is typically under 100 ns. These times apply when $R_L \leq 500\Omega$ and $C_O \leq 25$ pF.

Extra care must be taken in board layout since this is usually the dominant factor in satisfactory test results when measuring settling time. Short leads, 100 μF supply bypassing for low frequencies, and minimum scope lead length are all mandatory.

National Semiconductor

DAC0830/DAC0831/DAC0832 8-Bit μP Compatible, Double-Buffered D to A Converters

General Description

The DAC0830 is an advanced CMOS/Si-Cr 8-bit multiplying DAC designed to interface directly with the 8080, 8048, 8085, Z80®, and other popular microprocessors. A deposited silicon-chromium R-2R resistor ladder network divides the reference current and provides the circuit with excellent temperature tracking characteristics (0.05% of Full Scale Range maximum linearity error over temperature). The circuit uses CMOS current switches and control logic to achieve low power consumption and low output leakage current errors. Special circuitry provides TTL logic input voltage level compatibility.

Double buffering allows these DACs to output a voltage corresponding to one digital word while holding the next digital word. This permits the simultaneous updating of any number of DACs.

The DAC0830 series are the 8-bit members of a family of microprocessor-compatible DACs (MICRO-DAC™). For applications demanding higher resolution, the DAC1000 series (10-bits) and the DAC1208 and DAC1230 (12-bits) are available alternatives.

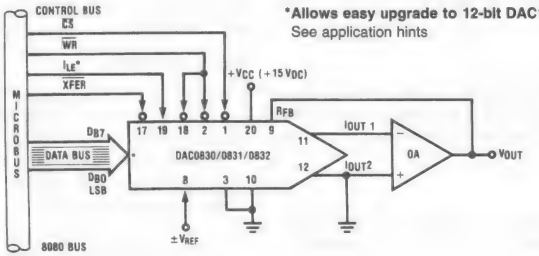
Features

- Double-buffered, single-buffered or flow-through digital data inputs
- Easy interchange and pin-compatible with 12-bit DAC1230 series
- Direct interface to all popular microprocessors
- Linearity specified with zero and full scale adjust only—NOT BEST STRAIGHT LINE FIT.
- Works with ±10V reference-full 4-quadrant multiplication
- Can be used in the voltage switching mode
- Logic inputs which meet TTL voltage level specs (1.4V logic threshold)
- Operates "STAND ALONE" (without μP) if desired
- Available in 20-pin small-outline or molded chip carrier package

Key Specifications

■ Current settling time	1 μs
■ Resolution	8 bits
■ Linearity (guaranteed over temp.)	8, 9, or 10 bits
■ Gain Tempco	0.0002% FS/°C
■ Low power dissipation	20 mW
■ Single power supply	5 to 15 V_{DC}

Typical Application



*Allows easy upgrade to 12-bit DAC1230, See application hints

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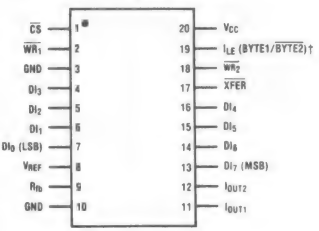
Connection Diagrams (Top Views)

Dual-In-Line and Small-Outline Packages



†This is necessary for the 12-bit DAC1230 series to permit interchanging from an 8-bit to a 12-bit DAC with **No PC board changes** and no software changes, See applications section.

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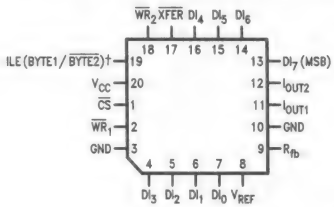
Molded Chip Carrier Package



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Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	17 V_{DC}
Voltage at Any Digital Input	V_{CC} to GND
Voltage at V_{REF} Input	±25V
Storage Temperature Range	−65°C to +150°C
Package Dissipation at T_A = 25°C (Note 3)	500 mW
DC Voltage Applied to I_{OUT1} or I_{OUT2} (Note 4)	−100 mV to V_{CC}
ESD Susceptability (Note 14)	800V
Lead Temperature (soldering, 10 sec.)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
Surface Mount Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

Operating Conditions

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
Part numbers with 'LCN' suffix	0°C to +70°C
Part numbers with 'LCWM' suffix	0°C to +70°C
Part numbers with 'LCV' suffix	0°C to +70°C
Part numbers with 'LCJ' suffix	−40°C to +85°C
Part numbers with 'LJ' suffix	−55°C to +125°C
Voltage at Any Digital Input	V_{CC} to GND

Electrical Characteristics

V_{REF} = 10.000 V_{DC} unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \le T_A \le T_{MAX}$.** For all other limits T_A = 25°C.

Parameter	Conditions	See Note	V_{CC} = 4.75 V_{DC} V_{CC} = 15.75 V_{DC}		V_{CC} = 5 V_{DC} ±5% V_{CC} = 12 V_{DC} ±5% to 15 V_{DC} ±5%	Limit Units
			Typ (Note 12)	Tested Limit (Note 5)	Design Limit (Note 6)	
CONVERTER CHARACTERISTICS						
Resolution			8	8	**8**	bits
Linearity Error Max	Zero and full scale adjusted $-10V \le V_{REF} \le +10V$	4, 8				
DAC0830LJ & LCJ				**0.05**	**0.05**	% FSR
DAC0832LJ & LCJ				**0.2**	**0.2**	% FSR
DAC0830LCN, LCWM & LCV				0.05	**0.05**	% FSR
DAC0831LCN				0.1	**0.1**	% FSR
DAC0832LCN, LCWM & LCV				0.2	**0.2**	% FSR
Differential Nonlinearity Max	Zero and full scale adjusted $-10V \le V_{REF} \le +10V$	4, 8				
DAC0830LJ & LCJ				**0.1**	**0.1**	% FSR
DAC0832LJ & LCJ				**0.4**	**0.4**	% FSR
DAC0830LCN, LCWM & LCV				0.1	**0.1**	% FSR
DAC0831LCN				0.2	**0.2**	% FSR
DAC0832LCN, LCWM & LCV				0.4	**0.4**	% FSR
Monotonicity	$-10V \le V_{REF} \le +10V$ LJ & LCJ	4		**8**	**8**	bits
	LCN, LCWM & LCV			8	**8**	bits
Gain Error Max	Using Internal R_{fb} $-10V \le V_{REF} \le +10V$	7	±0.2	±1	**±1**	% FS
Gain Error Tempco Max	Using internal R_{fb}		**0.0002**		**0.0006**	% FS/°C
Power Supply Rejection	All digital inputs latched high V_{CC} = 14.5V to 15.5V		0.0002	0.0025		% FSR/V
	11.5V to 12.5V		0.0006			
	4.5V to 5.5V		0.013	0.015		
Reference Input Max			**15**	**20**	**20**	kΩ
Reference Input Min			**15**	**10**	**10**	kΩ
Output Feedthrough Error	V_{REF} = 20 Vp-p, f = 100 kHz All data inputs latched low		3			mVp-p

4

Electrical Characteristics

$V_{REF} = 10.000\ V_{DC}$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \le T_A \le T_{MAX}$.** For all other limits $T_A = 25°C$. (Continued)

Parameter		Conditions	See Note	$V_{CC} = 4.75\ V_{DC}$ $V_{CC} = 15.75\ V_{DC}$		$V_{CC} = 5\ V_{DC} \pm 5\%$ $V_{CC} = 12\ V_{DC} \pm 5\%$ to $15\ V_{DC} \pm 5\%$	Limit Units
				Typ (Note 12)	Tested Limit (Note 5)	Design Limit (Note 6)	
CONVERTER CHARACTERISTICS (Continued)							
Output Leakage Current Max	I_{OUT1}	All data inputs latched low LJ & LCJ	10		**100**	**100**	nA
		All data inputs latched low LCN, LCWM & LCV			50	**100**	
	I_{OUT2}	All data inputs latched high LJ & LCJ			**100**	**100**	nA
		All data inputs latched high LCN, LCWM & LCV			50	**100**	
Output Capacitance	I_{OUT1} I_{OUT2}	All data inputs latched low		45 115			pF
	I_{OUT1} I_{OUT2}	All data inputs latched high		130 30			pF
DIGITAL AND DC CHARACTERISTICS							
Digital Input Voltages	Max	Logic Low LJ 4.75V			**0.6**		V_{DC}
		Logic Low LJ 15.75V			**0.8**		
		Logic Low LCJ 4.75V			**0.7**		
		Logic Low LCJ 15.75V			**0.8**		
		Logic Low LCN, LCWM, LCV			0.95	**0.8**	
	Min	Logic High LJ & LCJ			**2.0**	**2.0**	V_{DC}
		Logic High LCN, LCWM, LCV			1.9	**2.0**	
Digital Input Currents	Max	Digital inputs $<0.8V$ LJ & LCJ		−50	**−200**	**−200**	μA
		Digital inputs $<0.8V$ LCN, LCWM, LCV			−160	**−200**	μA
		Digital inputs $>2.0V$ LJ & LCJ		0.1	**+10**	**+10**	μA
		Digital inputs $>2.0V$ LCN, LCWM, LCV			+8	**+10**	
Supply Current Drain	Max	LJ & LCJ		1.2	**3.5**	**3.5**	mA
		LCN, LCWM, LCV			1.7	**2.0**	

Electrical Characteristics $V_{REF} = 10.000\ V_{DC}$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \le T_A \le T_{MAX}$.** For all other limits $T_A = 25°C$. (Continued)

Symbol	Parameter	Conditions	See Note	$V_{CC} = 15.75\ V_{DC}$		$V_{CC} = 12\ V_{DC} \pm 5\%$ to $15\ V_{DC} \pm 5\%$	$V_{CC} = 4.75\ V_{DC}$		$V_{CC} = 5\ V_{DC} \pm 5\%$	Limit Units
				Typ (Note 12)	Tested Limit (Note 5)	Design Limit (Note 6)	Typ (Note 12)	Tested Limit (Note 5)	Design Limit (Note 6)	
AC CHARACTERISTICS										
t_S	Current Setting Time	$V_{IL} = 0V$, $V_{IH} = 5V$		1.0			1.0			μs
t_W	Write and XFER Pulse Width Min	$V_{IL} = 0V$, $V_{IH} = 5V$	11 9	100	250 **320**	**320**	375	600 **900**	**900**	ns
t_{DS}	Data Setup Time Min	$V_{IL} = 0V$, $V_{IH} = 5V$	9	100	250 **320**	**320**	375	600 **900**	**900**	ns
t_{DH}	Data Hold Time Min	$V_{IL} = 0V$, $V_{IH} = 5V$	9		30 **30**			50 **50**		ns
t_{CS}	Control Setup Time Min	$V_{IL} = 0V$, $V_{IH} = 5V$	9	110	250 **320**	**320**	600	900 **1100**	**1100**	ns
t_{CH}	Control Hold Time Min	$V_{IL} = 0V$, $V_{IH} = 5V$	9	0	0 **0**	**10**	0	0 **0**		ns

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to GND, unless otherwise specified.

Note 3: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, θ_{JA}, and the ambient temperature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 125°C$ (plastic) or 150°C (ceramic), and the typical junction-to-ambient thermal resistance of the J package when board mounted is 80°C/W. For the N package, this number increases to 100°C/W and for the V package this number is 120°C/W.

Note 4: For current switching applications, both I_{OUT1} and I_{OUT2} must go to ground or the "Virtual Ground" of an operational amplifier. The linearity error is degraded by approximately $V_{OS} \div V_{REF}$. For example, if $V_{REF} = 10V$ then a 1 mV offset, V_{OS}, on I_{OUT1} or I_{OUT2} will introduce an additional 0.01% linearity error.

Note 5: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 6: Guaranteed, but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Note 7: Guaranteed at $V_{REF} = \pm 10\ V_{DC}$ and $V_{REF} = \pm 1\ V_{DC}$.

Note 8: The unit "FSR" stands for "Full Scale Range." "Linearity Error" and "Power Supply Rejection" specs are based on this unit to eliminate dependence on a particular V_{REF} value and to indicate the true performance of the part. The "Linearity Error" specification of the DAC0830 is "0.05% of FSR (MAX)". This guarantees that after performing a zero and full scale adjustment (see Sections 2.5 and 2.6), the plot of the 256 analog voltage outputs will each be within $0.05\% \times V_{REF}$ of a straight line which passes through zero and full scale.

Note 9: Boldface tested limits apply to the LJ and LCJ suffix parts only.

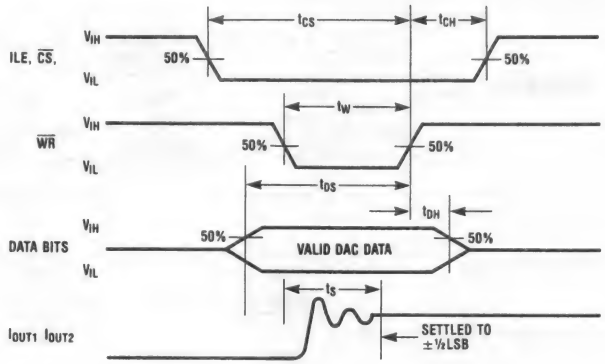
Note 10: A 100nA leakage current with $R_{fb} = 20k$ and $V_{REF} = 10V$ corresponds to a zero error of $(100 \times 10^{-9} \times 20 \times 10^3) \times 100/10$ which is 0.02% of FS.

Note 11: The entire write pulse must occur within the valid data interval for the specified t_W, t_{DS}, t_{DH}, and t_S to apply.

Note 12: Typicals are at 25°C and represent most likely parametric norm.

Note 13: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Switching Waveform



TL/H/5608–2

Definition of Package Pinouts

Control Signals (All control signals level actuated)

$\overline{CS}$: **Chip Select** (active low). The $\overline{CS}$ in combination with ILE will enable $\overline{WR_1}$.

ILE: **Input Latch Enable** (active high). The ILE in combination with $\overline{CS}$ enables $\overline{WR_1}$.

$\overline{WR_1}$: **Write 1.** The active low $\overline{WR_1}$ is used to load the digital input data bits (DI) into the input latch. The data in the input latch is latched when $\overline{WR_1}$ is high. To update the input latch—$\overline{CS}$ and $\overline{WR_1}$ must be low while ILE is high.

$\overline{WR_2}$: **Write 2** (active low). This signal, in combination with $\overline{XFER}$, causes the 8-bit data which is available in the input latch to transfer to the DAC register.

$\overline{XFER}$: **Transfer control signal** (active low). The $\overline{XFER}$ will enable $\overline{WR_2}$.

Other Pin Functions

DI_0-DI_7: **Digital Inputs.** DI_0 is the least significant bit (LSB) and DI_7 is the most significant bit (MSB).

I_{OUT1}: **DAC Current Output 1.** I_{OUT1} is a maximum for a digital code of all 1's in the DAC register, and is zero for all 0's in DAC register.

I_{OUT2}: **DAC Current Output 2.** I_{OUT2} is a constant minus I_{OUT1}, or $I_{OUT1} + I_{OUT2}$ = constant (I full scale for a fixed reference voltage).

R_{fb}: **Feedback Resistor.** The feedback resistor is provided on the IC chip for use as the shunt feedback resistor for the external op amp which is used to provide an output voltage for the DAC. This on-chip resistor should always be used (not an external resistor) since it matches the resistors which are used in the on-chip R-2R ladder and tracks these resistors over temperature.

V_{REF}: **Reference Voltage Input.** This input connects an external precision voltage source to the internal R-2R ladder. V_{REF} can be selected over the range of +10 to −10V. This is also the analog voltage input for a 4-quadrant multiplying DAC application.

V_{CC}: **Digital Supply Voltage.** This is the power supply pin for the part. V_{CC} can be from +5 to $+15V_{DC}$. Operation is optimum for $+15V_{DC}$.

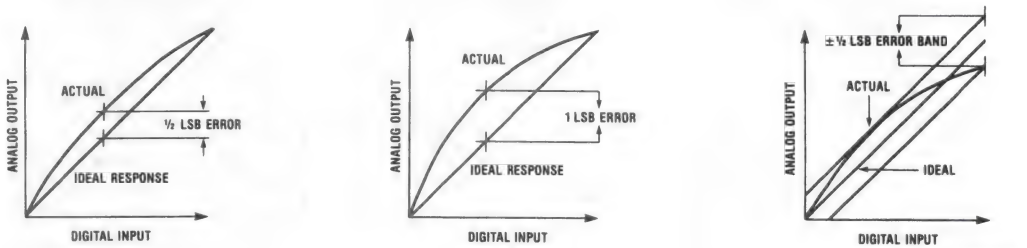
GND: The pin 10 voltage must be at the same ground potential as I_{OUT1} and I_{OUT2} for current switching applications. Any difference of potential (V_{OS} pin 10) will result in a linearity change of

$$\frac{V_{OS} \text{ pin } 10}{3V_{REF}}$$

For example, if V_{REF} = 10V and pin 10 is 9mV offset from I_{OUT1} and I_{OUT2} the linearity change will be 0.03%.

Pin 3 can be offset ±100mV with no linearity change, but the logic input threshold will shift.

Linearity Error



TL/H/5608–3

a) End point test after zero and fs adj.

b) Best straight line

c) Shifting fs adj. to pass best straight line test

Definition of Terms

Resolution: Resolution is directly related to the number of switches or bits within the DAC. For example, the DAC0830 has 2^8 or 256 steps and therefore has 8-bit resolution.

Linearity Error: Linearity Error is the maximum deviation from a *straight line passing through the endpoints of the DAC transfer characteristic.* It is measured after adjusting for zero and full-scale. Linearity error is a parameter intrinsic to the device and cannot be externally adjusted.

National's linearity "end point test" (a) and the "best straight line" test (b,c) used by other suppliers are illustrated above. The "end point test" greatly simplifies the adjustment procedure by eliminating the need for multiple iterations of checking the linearity and then adjusting full scale until the linearity is met. The "end point test" guarantees that linearity is met after a single full scale adjust. (One adjustment vs. multiple iterations of the adjustment.) The "end point test" uses a standard zero and F.S. adjustment procedure and is a much more stringent test for DAC linearity.

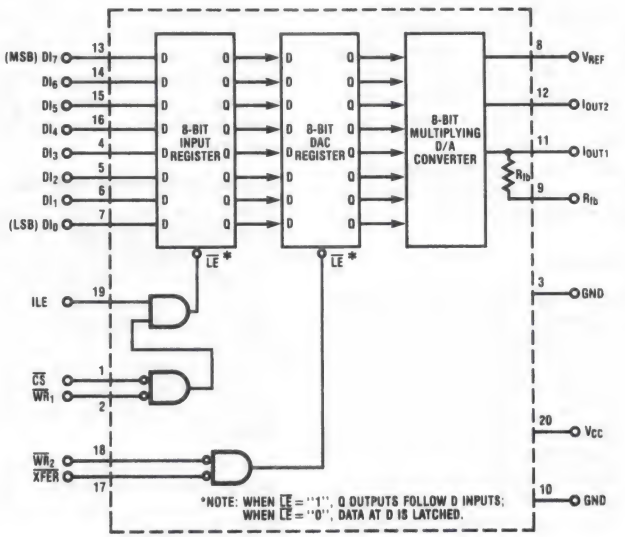
Power Supply Sensitivity: Power supply sensitivity is a measure of the effect of power supply changes on the DAC full-scale output.

Settling Time: Settling time is the time required from a code transition until the DAC output reaches within $\pm\frac{1}{2}$LSB of the final output value. Full-scale settling time requires a zero to full-scale or full-scale to zero output change.

Full-Scale Error: Full scale error is a measure of the output error between an ideal DAC and the actual device output. Ideally, for the DAC0830 series, full-scale is $V_{REF} - 1$LSB. For $V_{REF} = 10V$ and unipolar operation, $V_{FULL\text{-}SCALE} = 10.0000V - 39mV = 9.961V$. Full-scale error is adjustable to zero.

Differential Nonlinearity: The difference between any two consecutive codes in the transfer curve from the theoretical 1 LSB is differential nonlinearity.

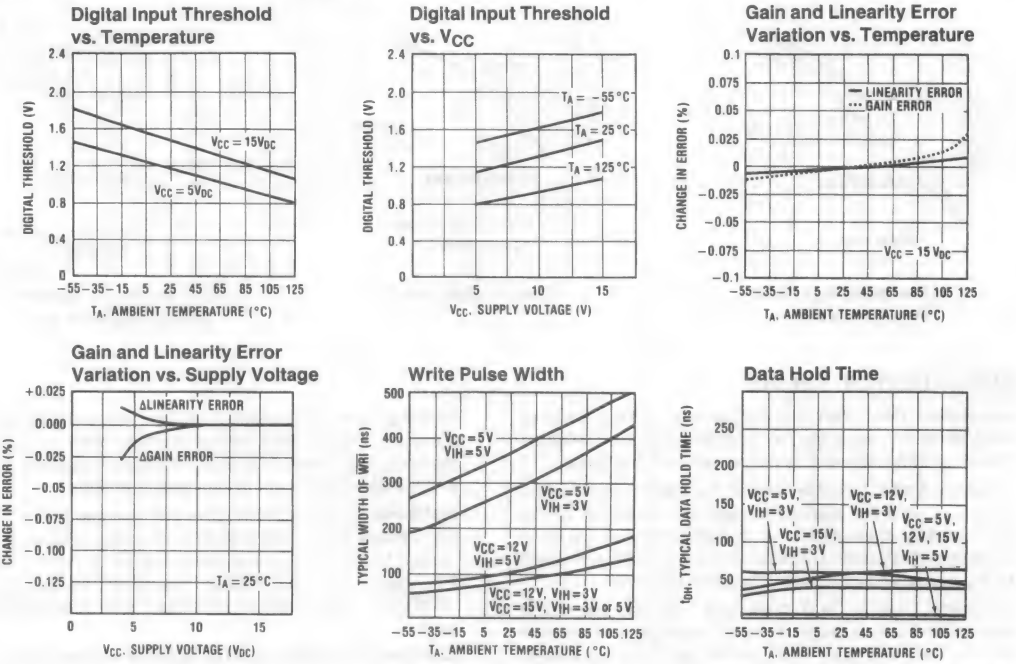
Monotonic: If the output of a DAC increases for increasing digital input code, then the DAC is monotonic. An 8-bit DAC which is monotonic to 8 bits simply means that increasing digital input codes will produce an increasing analog output.



TL/H/5608–4

FIGURE 1. DAC0830 Functional Diagram

Typical Performance Characteristics


Digital Input Threshold vs. Temperature
DIGITAL THRESHOLD (V)
$V_{CC} = 15V_{DC}$
$V_{CC} = 5V_{DC}$
T_A, AMBIENT TEMPERATURE (°C)
Digital Input Threshold vs. V_{CC}
$T_A = -55°C$
$T_A = 25°C$
$T_A = 125°C$
V_{CC}, SUPPLY VOLTAGE (V)
Gain and Linearity Error Variation vs. Temperature
LINEARITY ERROR
GAIN ERROR
CHANGE IN ERROR (%)
$V_{CC} = 15\ V_{DC}$
T_A, AMBIENT TEMPERATURE (°C)
Gain and Linearity Error Variation vs. Supply Voltage
ΔLINEARITY ERROR
ΔGAIN ERROR
$T_A = 25°C$
V_{CC}, SUPPLY VOLTAGE (V_{DC})
Write Pulse Width
TYPICAL WIDTH OF WR (ns)
$V_{CC} = 5V$, $V_{IH} = 5V$
$V_{CC} = 5V$, $V_{IH} = 3V$
$V_{CC} = 12V$, $V_{IH} = 5V$
$V_{CC} = 12V$, $V_{IH} = 3V$
$V_{CC} = 15V$, $V_{IH} = 3V$ or 5V
T_A, AMBIENT TEMPERATURE (°C)
Data Hold Time
t_{DH}, TYPICAL DATA HOLD TIME (ns)
$V_{CC} = 5V$, $V_{IH} = 3V$
$V_{CC} = 15V$, $V_{IH} = 3V$
$V_{CC} = 12V$, $V_{IH} = 3V$
$V_{CC} = 5V$, 12V, 15V, $V_{IH} = 5V$
T_A, AMBIENT TEMPERATURE (°C)

TL/H/5608–5

DAC0830 Series Application Hints

These DAC's are the industry's first microprocessor compatible, double-buffered 8-bit multiplying D to A converters. Double-buffering allows the utmost application flexibility from a digital control point of view. This 20-pin device is also pin for pin compatible (with one exception) with the DAC1230, a 12-bit MICRO-DAC. In the event that a system's analog output resolution and accuracy must be upgraded, substituting the DAC1230 can be easily accomplished. By tying address bit A_0 to the ILE pin, a two-byte μP write instruction (double precision) which automatically increments the address for the second byte write (starting with A_0 = "1") can be used. This allows either an 8-bit or the 12-bit part to be used with no hardware or software changes. For the simplest 8-bit application, this pin should be tied to V_{CC} (also see other uses in section 1.1).

Analog signal control versatility is provided by a precision R-2R ladder network which allows full 4-quadrant multiplication of a wide range bipolar reference voltage by an applied digital word.

1.0 DIGITAL CONSIDERATIONS

A most unique characteristic of these DAC's is that the 8-bit digital input byte is double-buffered. This means that the data must transfer through two independently controlled 8-bit latching registers before being applied to the R-2R ladder network to change the analog output. The addition of a second register allows two useful control features. First, any DAC in a system can simultaneously hold the current DAC data in one register (DAC register) and the next data word in the second register (input register) to allow fast updating of the DAC output on demand. Second, and probably more important, double-buffering allows any number of DAC's in a system to be updated to their new analog output levels simultaneously via a common strobe signal.

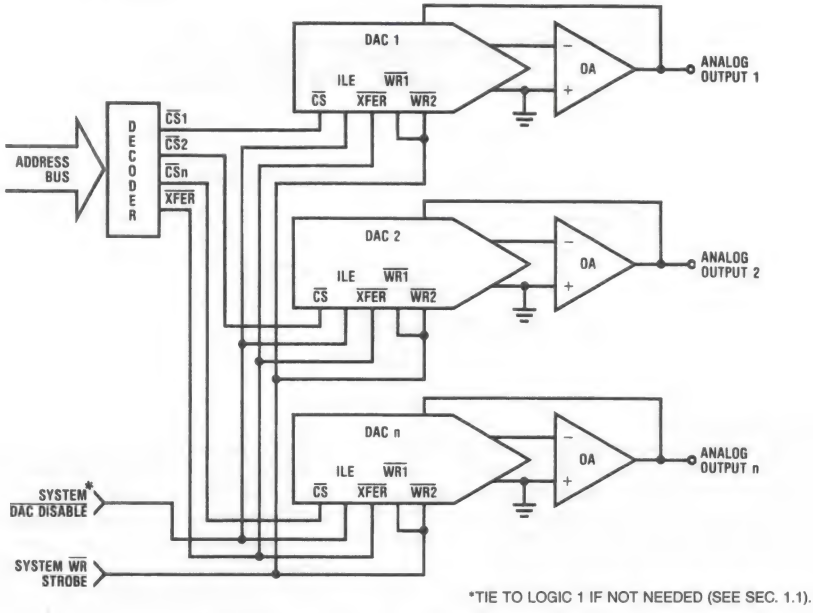
The timing requirements and logic level convention of the register control signals have been designed to minimize or eliminate external interfacing logic when applied to most popular microprocessors and development systems. It is easy to think of these converters as 8-bit "write-only" memory locations that provide an analog output quantity. All inputs to these DAC's meet TTL voltage level specs and can also be driven directly with high voltage CMOS logic in non-microprocessor based systems. To prevent damage to the chip from static discharge, all unused digital inputs should be tied to V_{CC} or ground. If any of the digital inputs are inadvertantly left floating, the DAC interprets the pin as a logic "1".

1.1 Double-Buffered Operation

Updating the analog output of these DAC's in a double-buffered manner is basically a two step or double write operation. In a microprocessor system two unique system addresses must be decoded, one for the input latch controlled by the $\overline{CS}$ pin and a second for the DAC latch which is controlled by the $\overline{XFER}$ line. If more than one DAC is being driven, *Figure 2*, the $\overline{CS}$ line of each DAC would typically be decoded individually, but all of the converters could share a common $\overline{XFER}$ address to allow simultaneous updating of any number of DAC's. The timing for this operation is shown, *Figure 3*.

It is important to note that the analog outputs that will change after a simultaneous transfer are those from the DAC's whose input register had been modified prior to the $\overline{XFER}$ command.

DAC0830 Series Application Hints (Continued)



*TIE TO LOGIC 1 IF NOT NEEDED (SEE SEC. 1.1).

FIGURE 2. Controlling Mutiple DACs

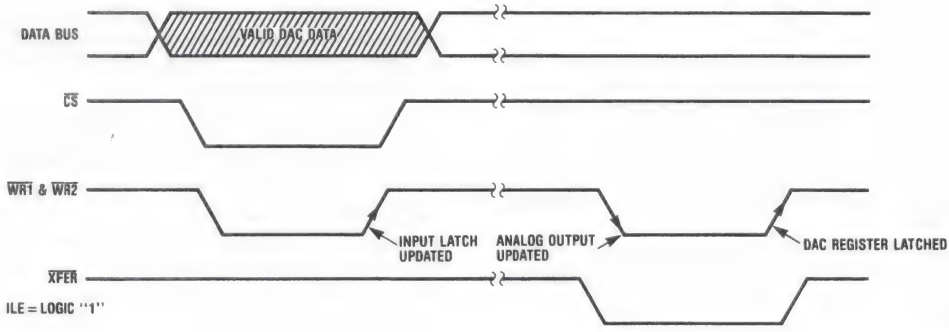


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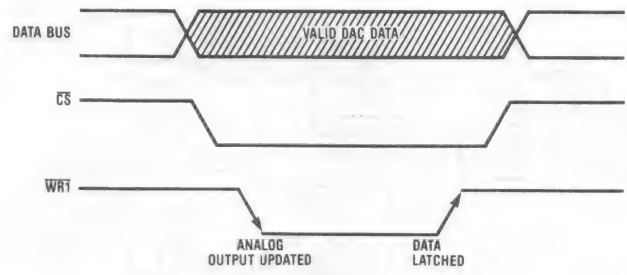
FIGURE 3

The ILE pin is an active high chip select which can be decoded from the address bus as a qualifier for the normal $\overline{CS}$ signal generated during a write operation. This can be used to provide a higher degree of decoding unique control signals for a particular DAC, and thereby create a more efficient addressing scheme.

Another useful application of the ILE pin of each DAC in a multiple DAC system is to tie these inputs together and use this as a control line that can effectively "freeze" the outputs of all the DAC's at their present value. Pulling this line low latches the input register and prevents new data from being written to the DAC. This can be particularly useful in multiprocessing systems to allow a processor other than the one controlling the DAC's to take over control of the data bus and control lines. If this second system were to use the same addresses as those decoded for DAC control (but for a different purpose) the ILE function would prevent the DAC's from being erroneously altered.

In a "Stand-Alone" system the control signals are generated by discrete logic. In this case double-buffering can be controlled by simply taking $\overline{CS}$ and $\overline{XFER}$ to a logic "0", ILE to a logic "1" and pulling $\overline{WR_1}$ low to load data to the input latch. Pulling $\overline{WR_2}$ low will then update the analog output. A logic "1" on either of these lines will prevent the changing of the analog output.

DAC0830 Series Application Hints (Continued)



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ILE = LOGIC "1"; $\overline{WR2}$ and $\overline{XFER}$ GROUNDED

FIGURE 4

1.2 Single-Buffered Operation

In a microprocessor controlled system where maximum data throughput to the DAC is of primary concern, or when only one DAC of several needs to be updated at a time, a single-buffered configuration can be used. One of the two internal registers allows the data to flow through and the other register will serve as the data latch.

Digital signal feedthrough (see Section 1.5) is minimized if the input register is used as the data latch. Timing for this mode is shown in *Figure 4*.

Single-buffering in a "stand-alone" system is achieved by strobing $\overline{WR_1}$ low to update the DAC with $\overline{CS}$, $\overline{WR_2}$ and $\overline{XFER}$ grounded and ILE tied high.

1.3 Flow-Through Operation

Though primarily designed to provide microprocessor interface compatibility, the MICRO-DAC's can easily be configured to allow the analog output to continuously reflect the state of an applied digital input. This is most useful in applications where the DAC is used in a continuous feedback control loop and is driven by a binary up-down counter, or in function generation circuits where a ROM is continuously providing DAC data.

Simply grounding $\overline{CS}$, $\overline{WR_1}$, $\overline{WR_2}$, and $\overline{XFER}$ and tying ILE high allows both internal registers to follow the applied digital inputs (flow-through) and directly affect the DAC analog output.

1.4 Control Signal Timing

When interfacing these MICRO-DAC to any microprocessor, there are two important time relationships that must be considered to insure proper operation. The first is the minimum $\overline{WR}$ strobe pulse width which is specified as 900 ns for all valid operating conditions of supply voltage and ambient temperature, but typically a pulse width of only 180ns is adequate if $V_{CC} = 15V_{DC}$. A second consideration is that the guaranteed minimum data hold time of 50ns should be met or erroneous data can be latched. This hold time is defined as the length of time data must be held valid on the digital inputs *after* a qualified (via $\overline{CS}$) $\overline{WR}$ strobe makes a low to high transition to latch the applied data.

If the controlling device or system does not inherently meet these timing specs the DAC can be treated as a slow memory or peripheral and utilize a technique to extend the write strobe. A simple extension of the write time, by adding a wait state, can simultaneously hold the write strobe active and data valid on the bus to satisfy the minimum $\overline{WR}$ pulsewidth. If this does not provide a sufficient data hold time at the end of the write cycle, a negative edge triggered oneshot can be included between the system write strobe and the $\overline{WR}$ pin of the DAC. This is illustrated in *Figure 5* for an exemplary system which provides a 250ns $\overline{WR}$ strobe time with a data hold time of less than 10ns.

The proper data set-up time prior to the latching edge (LO to HI transition) of the $\overline{WR}$ strobe, is insured if the $\overline{WR}$ pulsewidth is within spec and the data is valid on the bus for the duration of the DAC $\overline{WR}$ strobe.

1.5 Digital Signal Feedthrough

When data is latched in the internal registers, but the digital inputs are changing state, a narrow spike of current may flow out of the current output terminals. This spike is caused by the rapid switching of internal logic gates that are responding to the input changes.

There are several recommendations to minimize this effect. When latching data in the DAC, always use the input register as the latch. Second, reducing the V_{CC} supply for the DAC from +15V to +5V offers a factor of 5 improvement in the magnitude of the feedthrough, but at the expense of internal logic switching speed. Finally, increasing C_C (*Figure 8*) to a value consistent with the actual circuit bandwidth requirements can provide a substantial damping effect on any output spikes.

DAC0830 Series Application Hints (Continued)

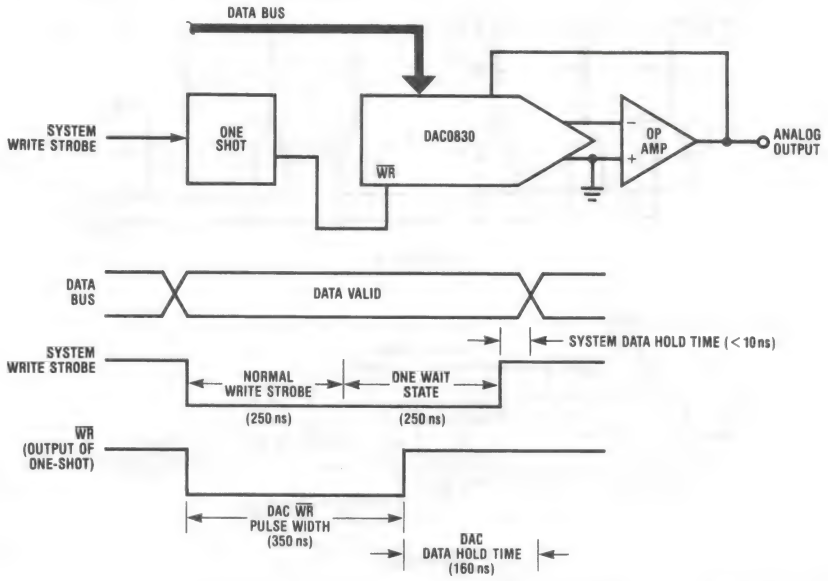

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FIGURE 5. Accommodating a High Speed System

2.0 ANALOG CONSIDERATIONS

The fundamental purpose of any D to A converter is to provide an accurate analog output quantity which is representative of the applied digital word. In the case of the DAC0830, the output, I_{OUT1}, is a current directly proportional to the product of the applied reference voltage and the digital input word. For application versatility, a second output, I_{OUT2}, is provided as a current directly proportional to the complement of the digital input. Basically:

$$I_{OUT1} = \frac{V_{REF}}{15\,k\Omega} \times \frac{\text{Digital Input}}{256};$$

$$I_{OUT2} = \frac{V_{REF}}{15\,k\Omega} \times \frac{255 - \text{Digital Input}}{256}$$

where the digital input is the decimal (base 10) equivalent of the applied 8-bit binary word (0 to 255), V_{REF} is the voltage at pin 8 and 15 kΩ is the nominal value of the internal resistance, R, of the R-2R ladder network (discussed in Section 2.1).

Several factors external to the DAC itself must be considered to maintain analog accuracy and are covered in subsequent sections.

2.1 The Current Switching R-2R Ladder

The analog circuitry, *Figure 6*, consists of a silicon-chromium (SiCr or Si-chrome) thin film R-2R ladder which is deposited on the surface oxide of the monolithic chip. As a result, there are no parasitic diode problems with the ladder (as there may be with diffused resistors) so the reference voltage, V_{REF}, can range −10V to +10V even if V_{CC} for the device is $5V_{DC}$.

The digital input code to the DAC simply controls the position of the SPDT current switches and steers the available ladder current to either I_{OUT1} or I_{OUT2} as determined by the logic input level ("1" or "0") respectively, as shown in *Figure 6*. The MOS switches operate in the current mode with a small voltage drop across them and can therefore switch currents of either polarity. This is the basis for the 4-quadrant multiplying feature of this DAC.

2.2 Basic Unipolar Output Voltage

To maintain linearity of output current with changes in the applied digital code, it is important that the voltages at both of the current output pins be as near ground potential ($0V_{DC}$) as possible. With $V_{REF} = +10V$ every millivolt appearing at either I_{OUT1} or I_{OUT2} will cause a 0.01% linearity error. In most applications this output current is converted to a voltage by using an op amp as shown in *Figure 7*.

The inverting input of the op amp is a "virtual ground" created by the feedback from its output through the internal 15 kΩ resistor, R_{fb}. All of the output current (determined by the digital input and the reference voltage) will flow through R_{fb} to the output of the amplifier. Two-quadrant operation can be obtained by reversing the polarity of V_{REF} thus causing I_{OUT1} to flow into the DAC and be sourced from the output of the amplifier. The output voltage, in either case, is always equal to $I_{OUT1} \times R_{fb}$ and is the opposite polarity of the reference voltage.

The reference can be either a stable DC voltage source or an AC signal anywhere in the range from −10V to +10V. The DAC can be thought of as a digitally controlled attenuator: the output voltage is always less than or equal to the applied reference voltage. The V_{REF} terminal of the device presents a nominal impedance of 15 kΩ to ground to external circuitry.

Always use the internal R_{fb} resistor to create an output voltage since this resistor matches (and tracks with temperature) the value of the resistors used to generate the output current (I_{OUT1}).

DAC0830 Series Application Hints (Continued)

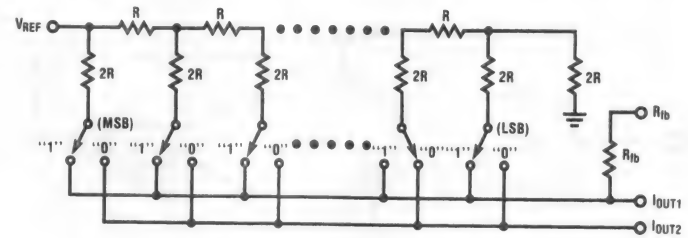

FIGURE 6

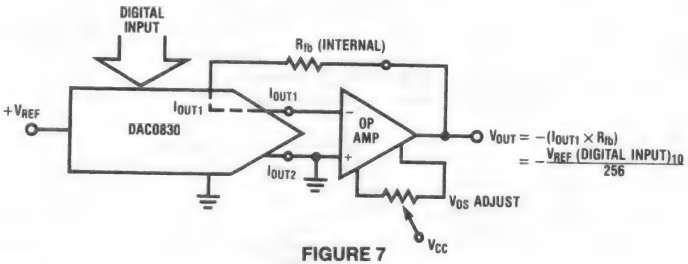

FIGURE 7

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2.3 Op Amp Considerations

The op amp used in *Figure 7* should have offset voltage nulling capability (See Section 2.5).

The selected op amp should have as low a value of input bias current as possible. The product of the bias current times the feedback resistance creates an output voltage error which can be significant in low reference voltage applications. BI-FET op amps are highly recommended for use with these DACs because of their very low input current.

Transient response and settling time of the op amp are important in fast data throughput applications. The largest stability problem is the feedback pole created by the feedback resistance, R_{fb}, and the output capacitance of the DAC. This appears from the op amp output to the (−) input and includes the stray capacitance at this node. Addition of a lead capacitance, C_C in *Figure 8*, greatly reduces overshoot and ringing at the output for a step change in DAC output current.

Finally, the output voltage swing of the amplifier must be greater than V_{REF} to allow reaching the full scale output voltage. Depending on the loading on the output of the amplifier and the available op amp supply voltages (only ±12 volts in many development systems), a reference voltage less than 10 volts may be necessary to obtain the full analog output voltage range.

2.4 Bipolar Output Voltage with a Fixed Reference

The addition of a second op amp to the previous circuitry can be used to generate a bipolar output voltage from a fixed reference voltage. This, in effect, gives sign significance to the MSB of the digital input word and allows two-quadrant multiplication of the reference voltage. The polarity of the reference can also be reversed to realize full 4-quadrant multiplication: $\pm V_{REF} \times \pm$Digital Code$= \pm V_{OUT}$. This circuit is shown in *Figure 9*.

This configuration features several improvements over existing circuits for bipolar outputs with other multiplying DACs. Only the offset voltage of amplifier 1 has to be nulled to preserve linearity of the DAC. The offset voltage error of the second op amp (although a constant output voltage error) has no effect on linearity. It should be nulled only if absolute output accuracy is required. Finally, the values of the resistors around the second amplifier do not have to match the internal DAC resistors, they need only to match and temperature track each other. A thin film 4-resistor network available from Beckman Instruments, Inc. (part no. 694-3-R10K-D) is ideally suited for this application. These resistors are matched to 0.1% and exhibit only 5 ppm/°C resistance tracking temperature coefficient. Two of the four available 10 kΩ resistors can be paralleled to form R in *Figure 9* and the other two can be used independently as the resistances labeled 2R.

2.5 Zero Adjustment

For accurate conversions, the input offset voltage of the output amplifier must always be nulled. Amplifier offset errors create an overall degradation of DAC linearity.

The fundamental purpose of zeroing is to make the voltage appearing at the DAC outputs as near $0V_{DC}$ as possible. This is accomplished for the typical DAC — op amp connection (*Figure 7*) by shorting out R_{fb}, the amplifier feedback resistor, and adjusting the V_{OS} nulling potentiometer of the op amp until the output reads zero volts. This is done, of course, with an applied digital code of all zeros if I_{OUT1} is driving the op amp (all one's for I_{OUT2}). The short around R_{fb} is then removed and the converter is zero adjusted.

DAC0830 Series Application Hints (Continued)

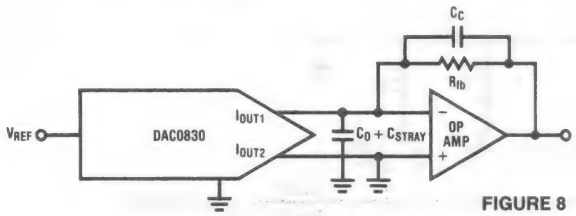
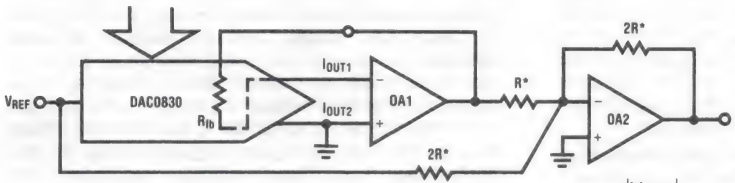


FIGURE 8

OP Amp	C_C	t_s (0 to Full Scale)
LF356	22 pF	4 μs
LF351	22 pF	5 μs
LF357*	10 pF	2 μs

*2.4 kΩ RESISTOR ADDED FROM − INPUT TO GROUND TO INSURE STABILITY



$$V_{OUT} = V_{REF} \frac{(\text{DIGITAL CODE} - 128)}{128}$$

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$$1\ \text{LSB} = \frac{|V_{REF}|}{128}$$

*THESE RESISTORS ARE AVAILABLE FROM BECKMAN INSTRUMENTS, INC. AS THEIR PART NO. 694-3-R10K-D

Input Code (MSB ..... LSB)	IDEAL V_{OUT} $+V_{REF}$	IDEAL V_{OUT} $-V_{REF}$
1 1 1 1 1 1 1 1	$V_{REF} - 1$ LSB	$-\lvert V_{REF}\rvert + 1$ LSB
1 1 0 0 0 0 0 0	$V_{REF}/2$	$-\lvert V_{REF}\rvert/2$
1 0 0 0 0 0 0 0	0	0
0 1 1 1 1 1 1 1	-1 LSB	$+1$ LSB
0 0 1 1 1 1 1 1	$-\frac{\lvert V_{REF}\rvert}{2} - 1$ LSB	$\frac{\lvert V_{REF}\rvert}{2} + 1$ LSB
0 0 0 0 0 0 0 0	$-\lvert V_{REF}\rvert$	$+\lvert V_{REF}\rvert$

FIGURE 9

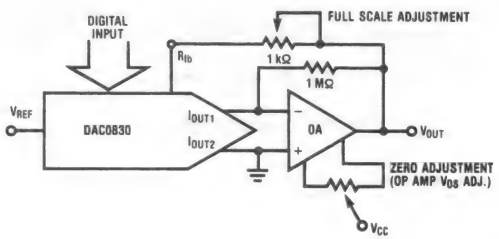
2.6 Full-Scale Adjustment

In the case where the matching of R_{fb} to the R value of the R-2R ladder (typically ±0.2%) is insufficient for full-scale accuracy in a particular application, the V_{REF} voltage can be adjusted or an external resistor and potentiometer can be added as shown in *Figure 10* to provide a full-scale adjustment.

The temperature coefficients of the resistors used for this adjustment are an important concern. To prevent degradation of the gain error temperature coefficient by the external resistors, their temperature coefficients ideally would have to match that of the internal DAC resistors, which is a highly impractical constraint. For the values shown in *Figure 10*, if the resistor and the potentiometer each had a temperature coefficient of ±100 ppm/°C maximum, the overall gain error temperature coefficent would be degraded a maximum of 0.0025%/°C for an adjustment pot setting of less than 3% of R_{fb}.

2.7 Using the DAC0830 in a Voltage Switching Configuration

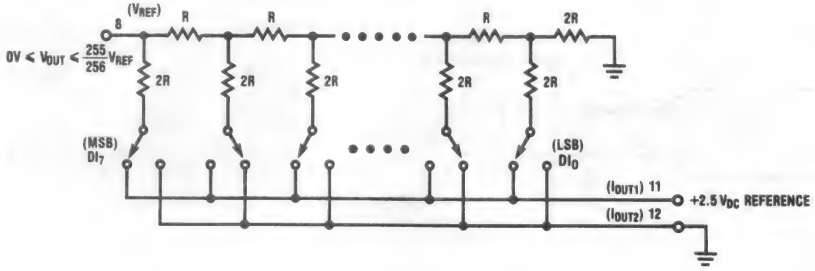
The R-2R ladder can also be operated as a voltage switching network. In this mode the ladder is used in an inverted manner from the standard current switching configuration. The reference voltage is connected to one of the current output terminals (I_{OUT1} for true binary digital control, I_{OUT2} is for complementary binary) and the output voltage is taken from the normal V_{REF} pin. The converter output is now a voltage in the range from 0V to 255/256 V_{REF} as a function of the applied digital code as shown in *Figure 11*.



TL/H/5608–11

FIGURE 10. Adding Full-Scale Adjustment

DAC0830 Series Application Hints (Continued)

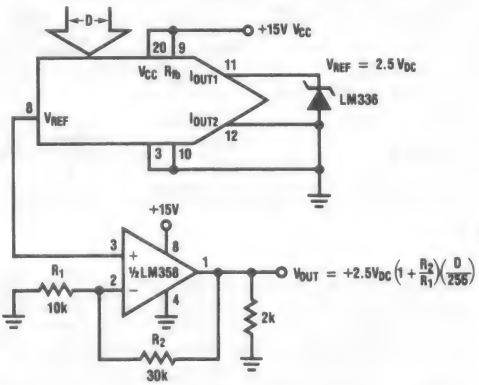


TL/H/5608–12

FIGURE 11. Voltage Mode Switching

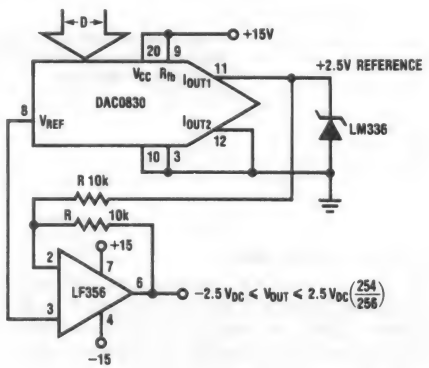
This configuration offers several useful application advantages. Since the output is a voltage, an external op amp is not necessarily required but the output impedance of the DAC is fairly high (equal to the specified reference input resistance of 10 kΩ to 20 kΩ) so an op amp may be used for buffering purposes. Some of the advantages of this mode are illustrated in *Figures 12, 13, 14* and *15*.

There are two important things to keep in mind when using this DAC in the voltage switching mode. The applied reference voltage must be positive since there are internal parasitic diodes from ground to the I_{OUT1} and I_{OUT2} terminals which would turn on if the applied reference went negative. There is also a dependence of conversion linearity and gain error on the voltage difference between V_{CC} and the voltage applied to the normal current output terminals. This is a result of the voltage drive requirements of the ladder switches. To ensure that all 8 switches turn on sufficiently (so as not to add significant resistance to any leg of the ladder and thereby introduce additional linearity and gain errors) it is recommended that the applied reference voltage be kept less than $+5V_{DC}$ and V_{CC} be at least 9V more positive than V_{REF}. These restrictions ensure less than 0.1% linearity and gain error change. *Figures 16, 17* and *18* characterize the effects of bringing V_{REF} and V_{CC} closer together as well as typical temperature performance of this voltage switching configuration.



- Voltage switching mode eliminates output signal inversion and therefore a need for a negative power supply.
- Zero code output voltage is limited by the low level output saturation voltage of the op amp. The 2 kΩ pull-down resistor helps to reduce this voltage.
- V_{OS} of the op amp has no effect on DAC linearity.

FIGURE 12. Single Supply DAC



TL/H/5608–13

- $V_{OUT} = 2.5V \left(\frac{D}{128} - 1\right)$
- Slewing and settling time for a full scale output change is ≈ 1.8 μs

FIGURE 13. Obtaining a Bipolar Output from a Fixed Reference with a Single Op Amp

DAC0830 Series Application Hints (Continued)

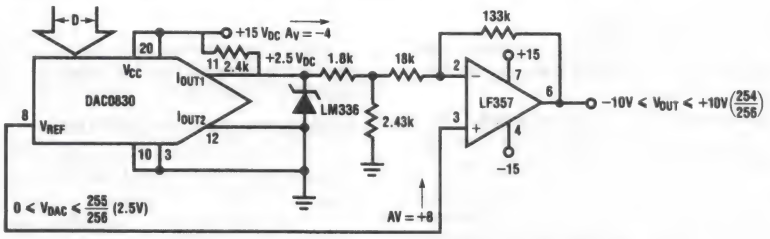
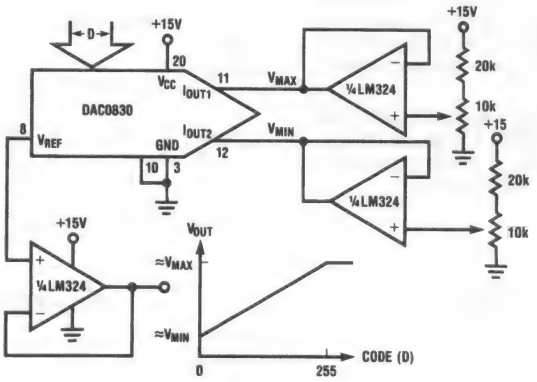


FIGURE 14. Bipolar Output with Increased Output Voltage Swing



TL/H/5608–14

- Only a single +15V supply required
- Non-interactive full-scale and zero code output adjustments
- V_{MAX} and V_{MIN} must be ≤ +5VDC and ≥0V.
- Incremental Output Step $= \frac{1}{256}(V_{MAX} - V_{MIN})$.
- $V_{OUT} = \frac{D}{256}(V_{MAX} - V_{MIN}) + \frac{255}{256}V_{MIN}$

FIGURE 15. Single Supply DAC with Level Shift and Span-Adjustable Output

Gain and Linearity Error Variation vs. Supply Voltage

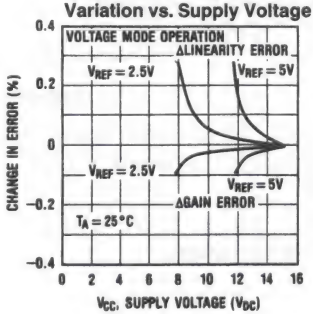


FIGURE 16

Gain and Linearity Error Variation vs. Reference Voltage

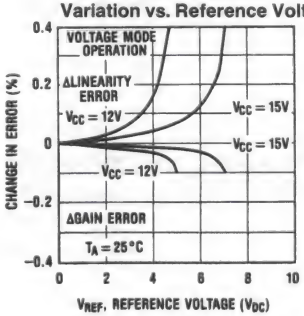
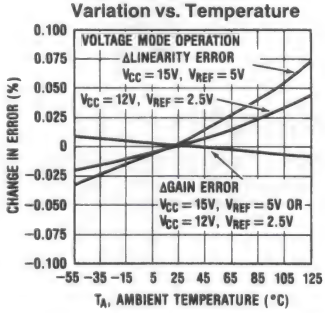


FIGURE 17

Note: For these curves, V_{REF} is the voltage applied to pin 11 (I_{OUT1}) with pin 12 (I_{OUT2}) grounded.

Gain and Linearity Error Variation vs. Temperature



TL/H/5608–15

FIGURE 18

DAC0830 Series Application Hints (Continued)

2.8 Miscellaneous Application Hints

These converters are CMOS products and reasonable care should be exercised in handling them to prevent catastrophic failures due to static discharge.

Conversion accuracy is only as good as the applied reference voltage so providing a stable source over time and temperature changes is an important factor to consider.

A "good" ground is most desirable. A single point ground distribution technique for analog signals and supply returns keeps other devices in a system from affecting the output of the DACs.

During power-up supply voltage sequencing, the −15V (or −12V) supply of the op amp may appear first. This will cause the output of the op amp to bias near the negative supply potential. No harm is done to the DAC, however, as the on-chip 15 kΩ feedback resistor sufficiently limits the current flow from I_{OUT1} when this lead is internally clamped to one diode drop below ground.

Careful circuit construction with minimization of lead lengths around the analog circuitry, is a primary concern. Good high frequency supply decoupling will aid in preventing inadvertant noise from appearing on the analog output.

Overall noise reduction and reference stability is of particular concern when using the higher accuracy versions, the DAC0830 and DAC0831, or their advantages are wasted.

3.0 GENERAL APPLICATION IDEAS

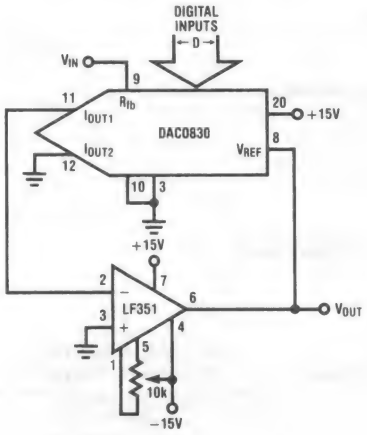
The connections for the control pins of the digital input registers are purposely omitted. Any of the control formats discussed in Section 1 of the accompanying text will work with any of the circuits shown. The method used depends on the overall system provisions and requirements.

The digital input code is referred to as D and represents the decimal equivalent value of the 8-bit binary input, for example:

Binary Input								
Pin 13 MSB							Pin 7 LSB	D Decimal Equivalent
1	1	1	1	1	1	1	1	255
1	0	0	0	0	0	0	0	128
0	0	0	1	0	0	0	0	16
0	0	0	0	0	0	1	0	2
0	0	0	0	0	0	0	0	0

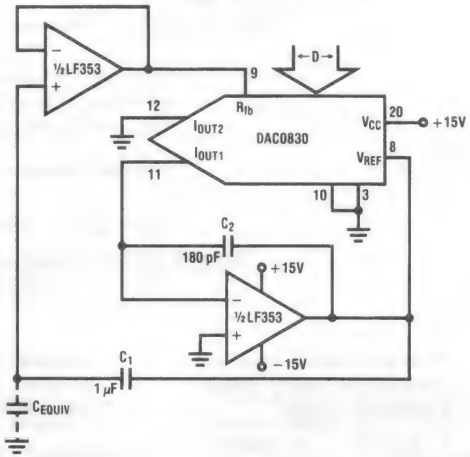
Applications

DAC Controlled Amplifier (Volume Control)



- $V_{OUT} = \frac{-V_{IN}(256)}{D}$
- When D = 0, the amplifier will go open loop and the output will saturate.
- Feedback impedance from the − input to the output varies from 15 kΩ to ∞ as the input code changes from full-scale to zero.

Capacitance Multiplier

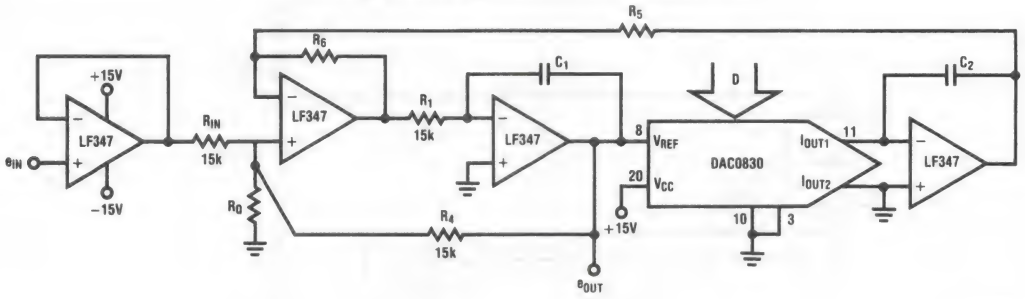


TL/H/5608–16

- $C_{EQUIV} = C_1\left(1 + \frac{256}{D}\right)$
- Maximum voltage across the equivalent capacitance is limited to $\frac{V_{O\,MAX}\ (\text{op amp})}{1 + \frac{256}{D}}$
- C_2 is used to improve settling time of op amp.

Applications (Continued)

Variable f_O, Variable Q_O, Constant BW Bandpass Filter



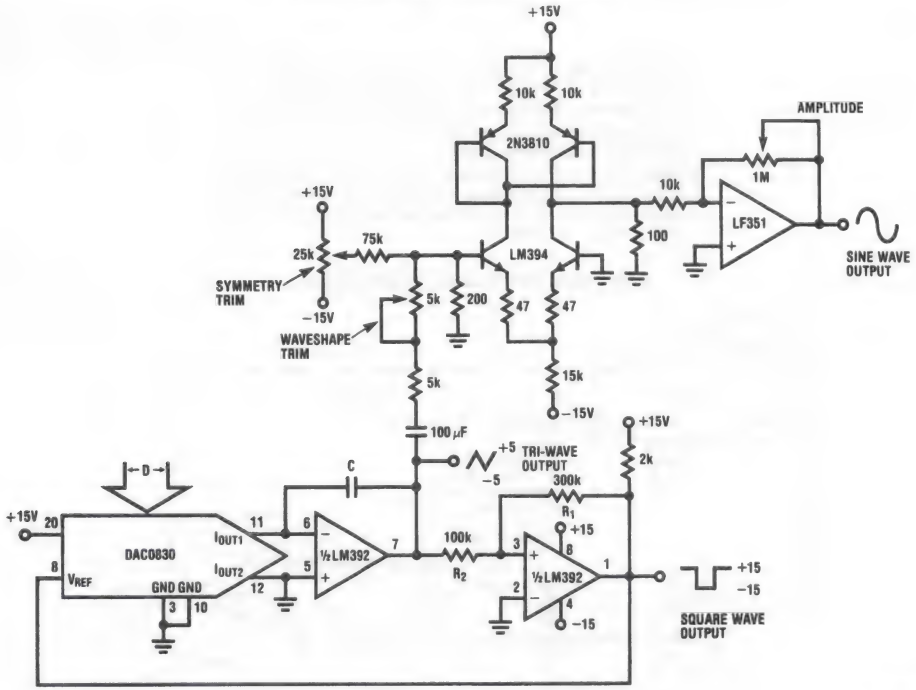
TL/H/5608–17

$$\bullet\ f_O = \frac{\sqrt{\frac{KD}{256}}}{2\pi R_1 C};\ Q_O = \sqrt{\frac{KD}{256}}\,\frac{(2R_Q + R_1)}{R_Q(K+1)};\ 3dbBW = \frac{R_Q(K+1)}{2\pi R_1 C(2R_Q + R1)}$$

where $C_1 = C_2 = C$; $K = \frac{R_6}{R_5}$ and $R_1 = R$ of DAC = 15k

- $H_O = 1$ for $R_{IN} = R_4 = R_1$
- Range of f_O and Q is ≈ 16 to 1 for circuit shown. The range can be extended to 255 to 1 by replacing R_1 with a second DAC0830 driven by the same digital input word.
- Maximum $f_O \times Q$ product should be ≤200 kHz.

DAC Controlled Function Generator

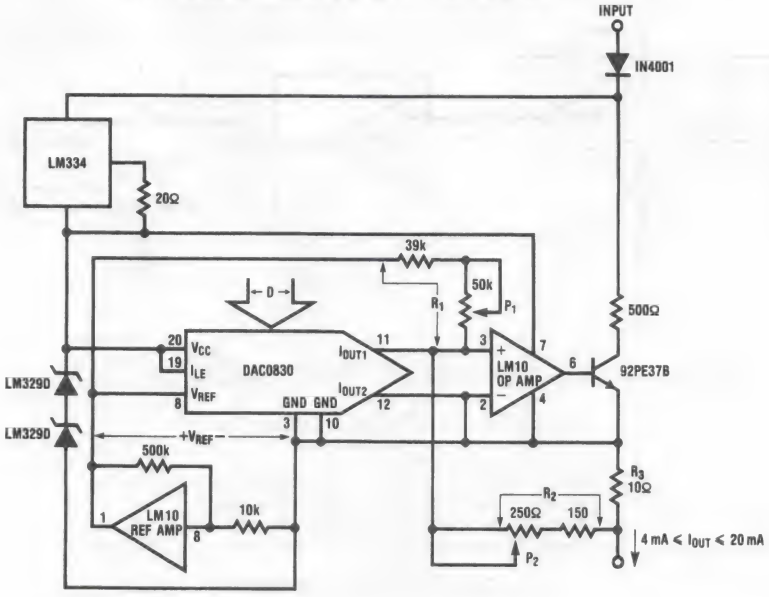


TL/H/5608–18

- DAC controls the frequency of sine, square, and triangle outputs.
- $f = \frac{D}{256(20k)C}$ for $V_{OMAX} = V_{0MIN}$ of square wave output and $R_1 = 3\,R_2$.
- 255 to 1 linear frequency range; oscillator stops with D = 0
- Trim symmetry and wave-shape for minimum sine wave distortion.

Applications (Continued)

Two Terminal Floating 4 to 20 mA Current Loop Controller

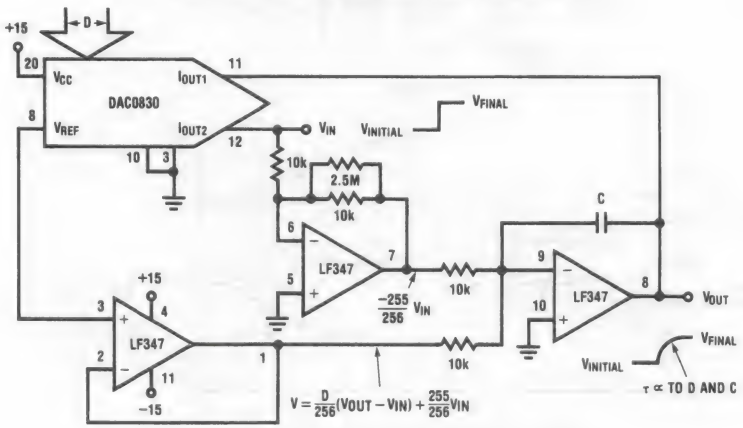


TL/H/5608–19

$$I_{OUT} = V_{REF}\left[\frac{1}{R_1} + \frac{D}{256\,R_{fb}}\right]\left[1 + \frac{R_2}{R_3}\right]$$

- DAC0830 linearly controls the current flow from the input terminal to the output terminal to be 4 mA (for D=0) to 19.94 mA (for D=255).
- Circuit operates with a terminal voltage differential of 16V to 55V.
- P_2 adjusts the magnitude of the output current and P_1 adjusts the zero to full scale range of output current.
- Digital inputs can be supplied from a processor using opto isolators on each input or the DAC latches can flow-through (connect control lines to pins 3 and 10 of the DAC) and the input data can be set by SPST toggle switches to ground (pins 3 and 10).

DAC Controlled Exponential Time Response



TL/H/5608–20

- Output responds exponentially to input changes and automatically stops when $V_{OUT}=V_{IN}$
- Output time constant is directly proportional to the DAC input code and capacitor C
- Input voltage must be positive (See section 2.7)

Ordering Information

Temperature Range		0°C to +70°			−40°C to +85°C	−55°C to +125°C
Non Linearity	0.05% FSR	DAC0830LCN	DAC0830LCM	DAC0830LCV	DAC0830LCJ	DAC0830LJ
	0.1% FSR	DAC0831LCN				
	0.2% FSR	DAC0832LCN	DAC0832LCM	DAC0832LCV	DAC0832LCJ	DAC0832LJ
Package Outline		N20A—Molded DIP	M20B Small Outline	V20A Chip Carrier	J20A—Ceramic DIP	

National Semiconductor

DAC1000/DAC1001/DAC1002/DAC1006/DAC1007/ DAC1008 μP Compatible, Double-Buffered D to A Converters

General Description

The DAC1000/1/2 and DAC1006/7/8 are advanced CMOS/Si-Cr 10-, 9- and 8-bit accurate multiplying DACs which are designed to interface directly with the 8080, 8048, 8085, Z-80 and other popular microprocessors. These DACs appear as a memory location or an I/O port to the μP and no interfacing logic is needed.

These devices, combined with an external amplifier and voltage reference, can be used as standard D/A converters; and they are very attractive for multiplying applications (such as digitally controlled gain blocks) since their linearity error is essentially independent of the voltage reference. They become equally attractive in audio signal processing equipment as audio gain controls or as programmable attenuators which marry high quality audio signal processing to digitally based systems under microprocessor control.

All of these DACs are double buffered. They can load all 10 bits or two 8-bit bytes and the data format can be either right justified or left justified. The analog section of these DACs is essentially the same as that of the DAC1020.

The DAC1000 series are the 10-bit members of a family of microprocessor-compatible DAC's (MICRO-DAC™'s). For applications requiring other resolutions, the DAC0830 series (8 bits) and the DAC1208 and DAC1230 (12 bits) are available alternatives.

Part #	Accuracy (bits)	Pin	Description
DAC1000	10	24	Has all logic features
DAC1001	9		
DAC1002	8		
DAC1006	10	20	For left-justified data
DAC1007	9		
DAC1008	8		

Features

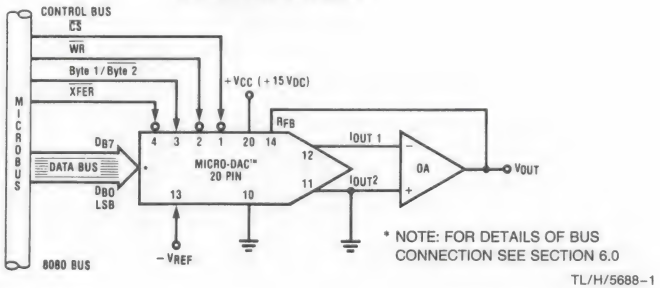
- Uses easy to adjust END POINT specs, NOT BEST STRAIGHT LINE FIT
- Low power consumption
- Direct interface to all popular microprocessors.
- Integrated thin film on CMOS structure
- Double-buffered, single-buffered or flow through digital data inputs.
- Loads two 8-bit bytes or a single 10-bit word.
- Logic inputs which meet TTL voltage level specs (1.4V logic threshold).
- Works with ±10V reference—full 4-quadrant multiplication.
- Operates STAND ALONE (without μP) if desired.
- Available in 0.3" standard 20-pin and 0.6" 24-pin package.
- Differential non-linearity selection available as special order.

Key Specifications

- Output Current Settling Time — 500 ns
- Resolution — 10 bits
- Linearity — 10, 9, and 8 bits (guaranteed over temp.)
- Gain Tempco — −0.0003% of FS/°C
- Low Power Dissipation (including ladder) — 20 mW
- Single Power Supply — 5 to 15 V_{DC}

Typical Application

DAC1006/1007/1008



TL/H/5688–1

Absolute Maximum Ratings (Notes 1 & 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	17 V_{DC}
Voltage at Any Digital Input	V_{CC} to GND
Voltage at V_{REF} Input	±25V
Storage Temperature Range	−65°C to +150°C
Package Dissipation at T_A = 25°C (Note 3)	500 mW
DC Voltage Applied to I_{OUT1} or I_{OUT2} (Note 4)	−100 mV to V_{CC}
ESD Susceptibility (Note 11)	800V
Lead Temp. (Soldering, 10 seconds)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C

Operating Ratings (Note 1)

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
Part numbers with 'LCN' suffix	0°C to 70°C
Part numbers with 'LCJ' suffix	−40°C to +85°C
Part numbers with 'LJ' suffix	−55°C to +125°C
Voltage at Any Digital Input	V_{CC} to GND

Electrical Characteristics

Tested at V_{CC} = 4.75 V_{DC} and 15.75 V_{DC}, T_A = 25°C, V_{REF} = 10.000 V_{DC} unless otherwise noted

Parameter	Conditions	See Note	V_{CC} = 12V_{DC} ±5% to 15V_{DC} ±5%			V_{CC} = 5V_{DC} ±5%			Units
			Min.	Typ.	Max.	Min.	Typ.	Max.	
Resolution					10			10	bits
Linearity Error	Endpoint adjust only	4,7							
	$T_{MIN} < T_A < T_{MAX}$	6							
	$-10V \le V_{REF} \le +10V$	5							
	DAC1000 and 1006				0.05			0.05	% of FSR
	DAC1001 and 1007				0.1			0.1	% of FSR
	DAC1002 and 1008				0.2			0.2	% of FSR
Differential Nonlinearity	Endpoint adjust only	4,7							
	$T_{MIN} < T_A < T_{MAX}$	6							
	$-10V \le V_{REF} \le +10V$	5							
	DAC1000 and 1006				0.1			0.1	% of FSR
	DAC1001 and 1007				0.2			0.2	% of FSR
	DAC1002 and 1008				0.4			0.4	% of FSR
Monotonicity	$T_{MIN} < T_A < T_{MAX}$	4,6							
	$-10V \le V_{REF} \le +10V$	5							
	DAC1000 and 1006		10			10			bits
	DAC1001 and 1007		9			9			bits
	DAC1002 and 1008		8			8			bits
Gain Error	Using internal R_{fb}								
	$-10V \le V_{REF} \le +10V$	5	−1.0	±0.3	1.0	−1.0	±0.3	1.0	% of FS
Gain Error Tempco	$T_{MIN} < T_A < T_{MAX}$	6							
	Using internal R_{fb}	9		−0.0003	−0.001		−0.0006	−0.002	% of FS/°C
Power Supply Rejection	All digital inputs latched high								
	V_{CC} = 14.5V to 15.5V			0.003	0.008				% FSR/V
	11.5V to 12.5V			0.004	0.010				% FSR/V
	4.75V to 5.25V						0.033	0.10	% FSR/V
Reference Input Resistance			10	15	20	10	15	20	kΩ
Output Feedthrough Error	V_{REF} = 20V_{p-p}, f = 100 kHz All data inputs latched low								
	D Package			130			130		mV_{p-p}
	N Package			90			90		mV_{p-p}
Output Capacitance I_{OUT1}	All data inputs latched low			60			60		pF
I_{OUT2}				250			250		pF
I_{OUT1}	All data inputs latched high			250			250		pF
I_{OUT2}				60			60		pF
Supply Current Drain	$T_{MIN} \le T_A \le T_{MAX}$	6		0.5	3.5		0.5	3.5	mA

Electrical Characteristics

Tested at V_{CC} = 4.75 V_{DC} and 15.75 V_{DC}, T_A = 25°C, V_{REF} = 10.000 V_{DC} unless otherwise noted (Continued)

Parameter	Conditions	See Note	V_{CC} = 12V_{DC} ± 5% to 15V_{DC} ± 5%			V_{CC} = 5V_{DC} ± 5%			Units
			Min.	Typ.	Max.	Min.	Typ.	Max.	
Output Leakage Current I_{OUT1}	$T_{MIN} \le T_A \le T_{MAX}$	6							
	All data inputs latched low	10			200			200	nA
I_{OUT2}	All data inputs latched high				200			200	nA
Digital Input Voltages	$T_{MIN} \le T_A \le T_{MAX}$	6							
	Low level								
	LJ suffix				0.8			0.6	V_{DC}
	LCJ, LCN suffix				0.8, 0.8			0.7, 0.8	V_{DC}
	High level (all parts)		2.0			2.0			V_{DC}
Digital Input Currents	$T_{MIN} \le T_A \le T_{MAX}$	6							
	Digital inputs <0.8V			−40	−150		−40	−150	μA_{DC}
	Digital inputs >2.0V			1.0	+10		1.0	+10	μA_{DC}
Current Settling Time t_S	V_{IL} = 0V, V_{IH} = 5V			500			500		ns
Write and $\overline{XFER}$ Pulse Width t_W	V_{IL} = 0V, V_{IH} = 5V, T_A = 25°C	8	150	60		320	200		ns
	$T_{MIN} \le T_A \le T_{MAX}$	9	320	100		500	250		ns
Data Set Up Time t_{DS}	V_{IL} = 0V, V_{IH} = 5V, T_A = 25°C	9	150	80		320	170		ns
	$T_{MIN} \le T_A \le T_{MAX}$		320	120		500	250		ns
Data Hold Time t_{DH}	V_{IL} = OV, V_{IH} = 5V T_A = 25°C	9	200	100		320	220		ns
	$T_{MIN} \le T_A \le T_{MAX}$		250	120		500	320		ns
Control Set Up Time t_{CS}	V_{IL} = 0V, V_{IL} = 5V, T_A = 25°C	9	150	60		320	180		ns
	$T_{MIN} \le T_A \le T_{MAX}$		320	100		500	260		ns
Control Hold Time t_{CH}	V_{IL} = 0V, V_{IH} = 5V, T_A = 25°C	9	10	0		10	0		ns
	$T_{MIN} \le T_A \le T_{MAX}$		10	0		10	0		ns

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to GND, unless otherwise specified.

Note 3: This 500 mW specification applies for all packages. The low intrinsic power dissipation of this part (and the fact that there is no way to significantly modify the power dissipation) removes concern for heat sinking.

Note 4: For current switching applications, both I_{OUT1} and I_{OUT2} must go to ground or the "Virtual Ground" of an operational amplifier. The linearity error is degraded by approximately $V_{OS} \div V_{REF}$. For example, if V_{REF} = 10V then a 1 mV offset, V_{OS}, on I_{OUT1} or I_{OUT2} will introduce an additional 0.01% linearity error.

Note 5: Guaranteed at V_{REF} = ±10 V_{DC} and V_{REF} = ±1 V_{DC}.

Note 6: T_{MIN} = 0°C and T_{MAX} = 70°C for "LCN" suffix parts.

T_{MIN} = −40°C and T_{MAX} = 85°C for "LCJ" suffix parts.

T_{MIN} = 55°C and T_{MAX} = 125°C for "LJ" suffix parts.

Note 7: The unit "FSR" stands for "Full Scale Range." "Linearity Error" and "Power Supply Rejection" specs are based on this unit to eliminate dependence on a particular V_{REF} value and to indicate the true performance of the part. The "Linearity Error" specification of the DAC1000 is "0.05% of FSR (MAX)." This guarantees that after performing a zero and full scale adjustment (See Sections 2.5 and 2.6), the plot of the 1024 analog voltage outputs will each be within 0.05% × V_{REF} of a straight line which passes through zero and full scale.

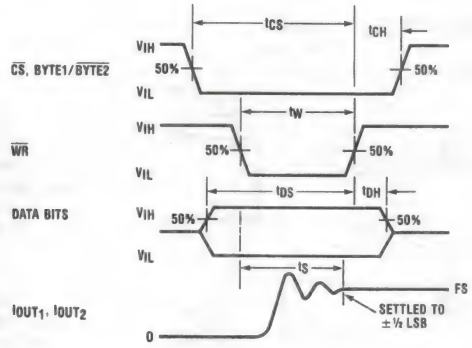
Note 8: This specification implies that all parts are guaranteed to operate with a write pulse or transfer pulse width (t_W) of 320 ns. A typical part will operate with t_W of only 100 ns. The entire write pulse must occur within the valid data interval for the specified t_W, t_{DS}, t_{DH}, and t_S to apply.

Note 9: Guaranteed by design but not tested.

Note 10: A 200 nA leakage current with R_{fb} = 20K and V_{REF} = 10V corresponds to a zero error of $(200 \times 10^{-9} \times 20 \times 10^{3}) \times 100 \div 10$ which is 0.04% of FS.

Note 11: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

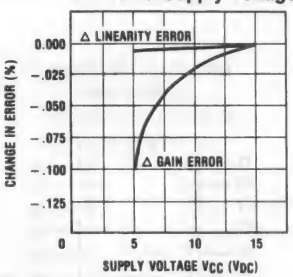
Switching Waveforms



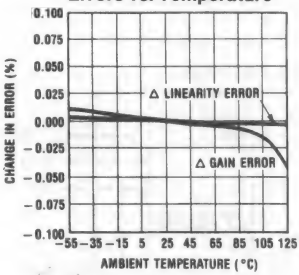
TL/H/5688–2

Typical Performance Characteristics

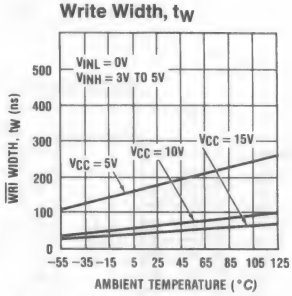
Errors vs. Supply Voltage



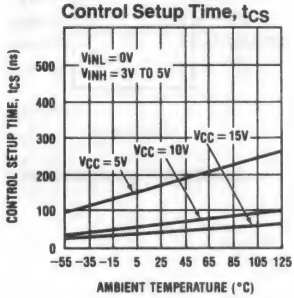
Errors vs. Temperature



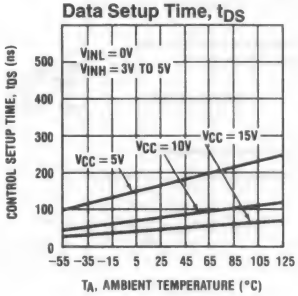
Write Width, t_W



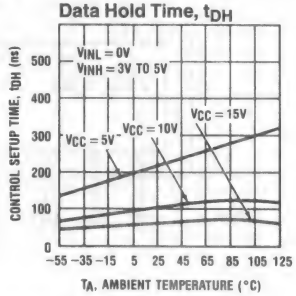
Control Setup Time, t_{CS}



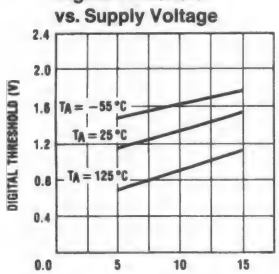
Data Setup Time, t_{DS}



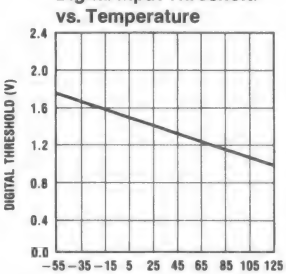
Data Hold Time, t_{DH}



Digital Threshold vs. Supply Voltage



Digital Input Threshold vs. Temperature

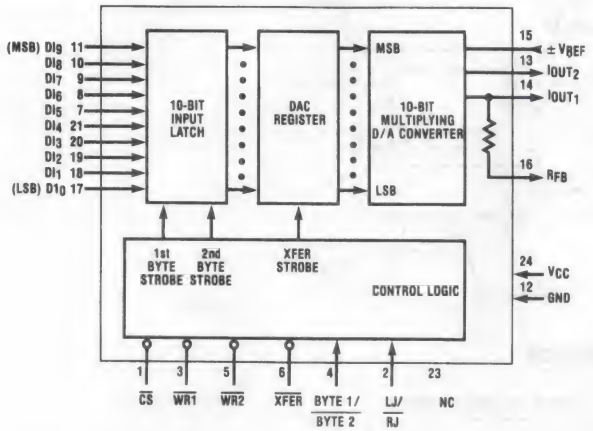


TL/H/5688–3

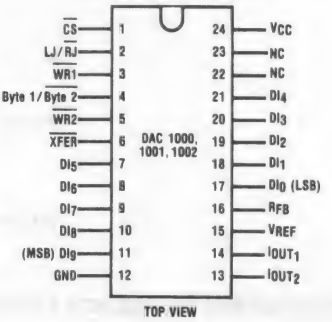
4

Block and Connection Diagrams

DAC1000/1001/1002 (24-Pin Parts)



DAC1000/1001/1002
(24-Pin Parts)
Dual-In-Line Package



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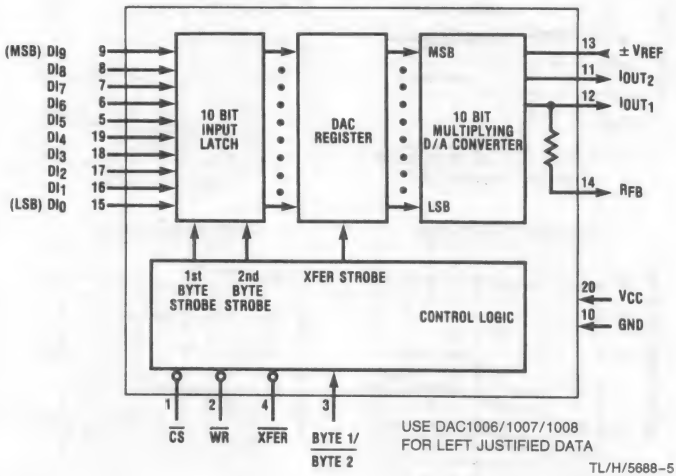
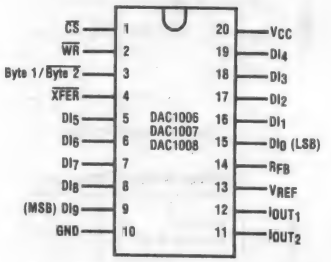
DAC1006/1007/1008 (20-Pin Parts)



USE DAC1006/1007/1008
FOR LEFT JUSTIFIED DATA

TL/H/5688–5

DAC1006/1007/1008
(20-Pin Parts)
Dual-In-Line Package

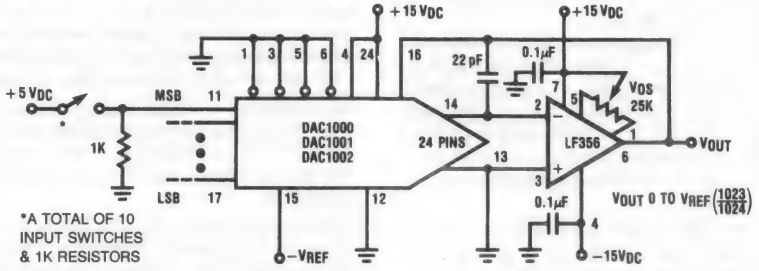


TL/H/5688–28

Top View

See Ordering Information

DAC1000/1001/1002—Simple Hookup for a "Quick Look"

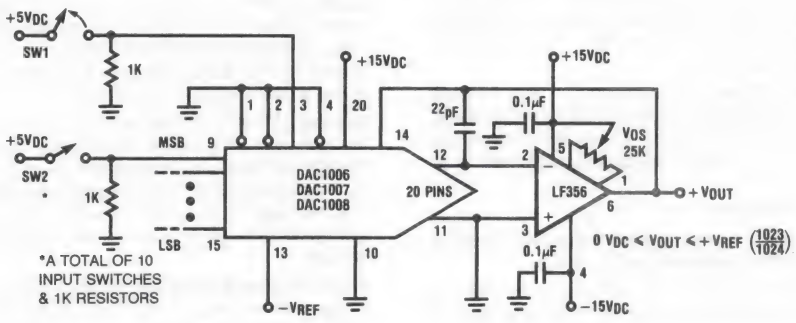


TL/H/5688–6

Notes:

1. For $V_{REF} = -10.240\ V_{DC}$ the output voltage steps are approximately 10 mV each.
2. Operation is set up for flow through—no latching of digital input data.
3. Single point ground is strongly recommended.

DAC1006/1007/1008—Simple Hookup for a "Quick Look"



TL/H/5688–7

Notes:

1. For $V_{REF} = -10.240\ V_{DC}$ the output voltage steps are approximately 10 mV each.
2. SW1 is a normally closed switch. While SW1 is closed, the DAC register is latched and new data can be loaded into the input latch via the 10 SW2 switches.

When SW1 is momentarily opened the new data is transferred from the input latch to the DAC register and is latched when SW1 again closes.

1.0 DEFINITION OF PACKAGE PINOUTS

1.1 Control Signals (All control signals are level actuated.)

$\overline{CS}$: Chip Select — active low, it will enable $\overline{WR}$ (DAC1003–1008) or WR_1 (DAC1000–1002).

$\overline{WR}$ or $\overline{WR_1}$: Write — The active low $\overline{WR}$ (or $\overline{WR_1}$ — DAC1000-1002) is used to load the digital data bits (DI) into the input latch. The data in the input latch is latched when $\overline{WR}$ (or $\overline{WR_1}$) is high. The 10-bit input latch is split into two latches; one holds 8 bits and the other holds 2 bits. The Byte1/$\overline{Byte2}$ control pin is used to select both input latches when Byte1/$\overline{Byte2}$ = 1 or to overwrite the 2-bit input latch when in the low state.

$\overline{WR_2}$: Extra Write (DAC1000–1002) — The active low $\overline{WR_2}$ is used to load the data from the input latch to the DAC register while $\overline{XFER}$ is low. The data in the DAC register is latched when WR_2 is high.

Byte1/$\overline{Byte2}$: Byte Sequence Control — When this control is high, all ten locations of the input latch are enabled. When low, only two locations of the input latch are enabled and these two locations are overwritten on the second byte write. On the DAC1006, 1007, and 1008, the Byte1/$\overline{Byte2}$ must be low to transfer the 10-bit data in the input latch to the DAC register.

$\overline{XFER}$: Transfer Control Signal, active low — This signal, in combination with others, is used to transfer the 10-bit data which is available in the input latch to the DAC register — see timing diagrams.

LJ/$\overline{RJ}$: Left Justify/Right Justify (DAC1000–1002) — When LJ/$\overline{RJ}$ is high the part is set up for left justified (fractional) data format. (DAC1006–1008 have this done internally.) When LJ/$\overline{RJ}$ is low, the part is set up for right justified (integer) data.

1.2 Other Pin Functions

DI_i (i = 0 to 9): Digital Inputs — DI_0 is the least significant bit (LSB) and DI_9 is the most significant bit (MSB).

I_{OUT1}: DAC Current Output 1 — I_{OUT1} is a maximum for a digital input code of all 1s and is zero for a digital input code of all 0s.

I_{OUT2}: DAC Current Output 2 — I_{OUT2} is a constant minus I_{OUT1}, or

$$I_{OUT1} + I_{OUT2} = \frac{1023\ V_{REF}}{1024\ R}$$

where R ≅ 15 kΩ.

R_{FB}: Feedback Resistor — This is provided on the IC chip for use as the shunt feedback resistor when an external op amp is used to provide an output voltage for the DAC. This on-chip resistor should always be used (not an external resistor) because it matches the resistors used in the on-chip R-2R ladder and tracks these resistors over temperature.

V_{REF}: Reference Voltage Input — This is the connection for the external precision voltage source which drives the R-2R ladder. V_{REF} can range from −10 to +10 volts. This is also the analog voltage input for a 4-quadrant multiplying DAC application.

V_{CC}: Digital Supply Voltage — This is the power supply pin for the part. V_{CC} can be from +5 to +15 V_{DC}. Operation is optimum for +15V. The input threshold voltages are nearly independent of V_{CC}. (See Typical Performance Characteristics and Description in Section 3.0, T^2L compatible logic inputs.)

GND: Ground — the ground pin for the part.

1.3 Definition of Terms

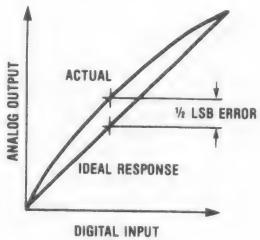
Resolution: Resolution is directly related to the number of switches or bits within the DAC. For example, the DAC1000 has 2^{10} or 1024 steps and therefore has 10-bit resolution.

Linearity Error: Linearity error is the maximum deviation from a *straight line passing through the endpoints of the DAC transfer characteristic.* It is measured after adjusting for zero and full-scale. Linearity error is a parameter intrinsic to the device and cannot be externally adjusted.

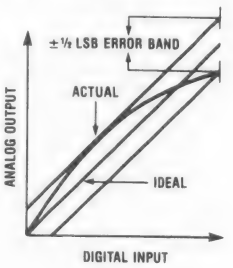
National's linearity test (a) and the "best straight line" test (b) used by other suppliers are illustrated below. The "best straight line" requires a special zero and FS adjustment for each part, which is almost impossible for user to determine. The "end point test" uses a standard zero and FS adjustment procedure and is a much more stringent test for DAC linearity.

Power Supply Sensitivity: Power supply sensitivity is a measure of the effect of power supply changes on the DAC full-scale output (which is the worst case).

a. End Point Test After Zero and FS Adj.



b. Best Straight Line



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Settling Time: Settling time is the time required from a code transition until the DAC output reaches within $\pm\frac{1}{2}$ LSB of the final output value. Full-scale settling time requires a zero to full-scale or full-scale to zero output change.

Full-Scale Error: Full scale error is a measure of the output error between an ideal DAC and the actual device output. Ideally, for the DAC1000 series, full-scale is $V_{REF}-1$ LSB. For $V_{REF}=-10V$ and unipolar operation, $V_{FULL\text{-}SCALE}=10.0000V-9.8mV=9.9902V$. Full-scale error is adjustable to zero.

Monotonicity: If the output of a DAC increases for increasing digital input code, then the DAC is monotonic. A 10-bit DAC with 10-bit monotonicity will produce an increasing analog output when all 10 digital inputs are exercised. A 10-bit DAC with 9-bit monotonicity will be monotonic when only the most significant 9 bits are exercised. Similarly, 8-bit monotonicity is guaranteed when only the most significant 8 bits are exercised.

2.0 DOUBLE BUFFERING

These DACs are double-buffered, microprocessor compatible versions of the DAC1020 10-bit multiplying DAC. The addition of the buffers for the digital input data not only allows for storage of this data, but also provides a way to assemble the 10-bit input data word from two write cycles when using an 8-bit data bus. Thus, the next data update for the DAC output can be made with the complete new set of 10-bit data. Further, the double buffering allows many DACs in a system to store current data and also the next data. The updating of the new data for each DAC is also not time critical. When all DACs are updated, a common strobe signal can then be used to cause all DACs to switch to their new analog output levels.

3.0 TTL COMPATIBLE LOGIC INPUTS

To guarantee TTL voltage compatibility of the logic inputs, a novel bipolar (NPN) regulator circuit is used. This makes the input logic thresholds equal to the forward drop of two diodes (and also matches the temperature variation) as occurs naturally in TTL. The basic circuit is shown in *Figure 1*. A curve of digital input threshold as a function of power supply voltage is shown in the Typical Performance Characteristics section.

4.0 APPLICATION HINTS

The DC stability of the V_{REF} source is the most important factor to maintain accuracy of the DAC over time and temperature changes. A good single point ground for the analog signals is next in importance.

These MICRO-DAC converters are CMOS products and reasonable care should be exercised in handling them prior to final mounting on a PC board. The digital inputs are protected, but permanent damage may occur if the part is subjected to high electrostatic fields. Store unused parts in conductive foam or anti-static rails.

4.1 Power Supply Sequencing & Decoupling

Some IC amplifiers draw excessive current from the Analog inputs to V− when the supplies are first turned on. To prevent damage to the DAC — an external Schottky diode connected from I_{OUT1} or I_{OUT2} to ground may be required to prevent destructive currents in I_{OUT1} or I_{OUT2}. If an LM741 or LF356 is used — these diodes are not required.

The standard power supply decoupling capacitors which are used for the op amp are adequate for the DAC.

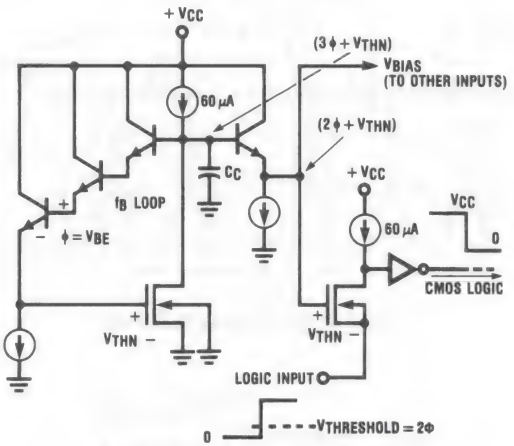

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FIGURE 1. Basic Logic Threshold Loop

4.2 Op Amp Bias Current & Input Leads

The op amp bias current (I_B) CAN CAUSE DC ERRORS. BI-FET™ op amps have very low bias current, and therefore the error introduced is negligible. BI-FET op amps are strongly recommended for these DACs.

The distance from the I_{OUT1} pin of the DAC to the inverting input of the op amp should be kept as short as possible to prevent inadvertent noise pickup.

5.0 ANALOG APPLICATIONS

The analog section of these DACs uses an R-2R ladder which can be operated both in the current switching mode and in the voltage switching mode.

The major product changes (compared with the DAC1020) have been made in the digital functioning of the DAC. The analog functioning is reviewed here for completeness. For additional analog applications, such as multipliers, attenuators, digitally controlled amplifiers and low frequency sine wave oscillators, refer to the DAC1020 data sheet. Some basic circuit ideas are presented in this section in addition to complete applications circuits.

5.1 Operation in Current Switching Mode

The analog circuitry, *Figure 2*, consists of a silicon-chromium (Si-Cr) thin film R-2R ladder which is deposited on the surface oxide of the monolithic chip. As a result, there is no parasitic diode connected to the V_{REF} pin as would exist if diffused resistors were used. The reference voltage input (V_{REF}) can therefore range from −10V to +10V.

The digital input code to the DAC simply controls the position of the SPDT current switches, SW0 to SW9. A logical 1 digital input causes the current switch to steer the available ladder current to the I_{OUT1} output pin. These MOS switches operate in the current mode with a small voltage drop across them and can therefore switch currents of either polarity. This is the basis for the 4-quadrant multiplying feature of this DAC.

5.1.1 Providing a Unipolar Output Voltage with the DAC in the Current Switching Mode

A voltage output is provided by making use of an external op amp as a current-to-voltage converter. The idea is to use the internal feedback resistor, R_{FB}, from the output of the op amp to the inverting (−) input. Now, when current is entered at this inverting input, the feedback action of the op amp keeps that input at ground potential. This causes the applied input current to be diverted to the feedback resistor. The output voltage of the op amp is forced to a voltage given by:

$$V_{OUT} = -(I_{OUT1} \times R_{FB})$$

Notice that the sign of the output voltage depends on the direction of current flow through the feedback resistor.

In current switching mode applications, both current output pins (I_{OUT1} and I_{OUT2}) should be operated at 0 V_{DC}. This is accomplished as shown in *Figure 3*. The capacitor, C_C, is used to compensate for the output capacitance of the DAC and the input capacitance of the op amp. The required feedback resistor, R_{FB}, is available on the chip (one end is internally tied to I_{OUT1}) and must be used since an external resistor will not provide the needed matching and temperature tracking. This circuit can therefore be simplified as

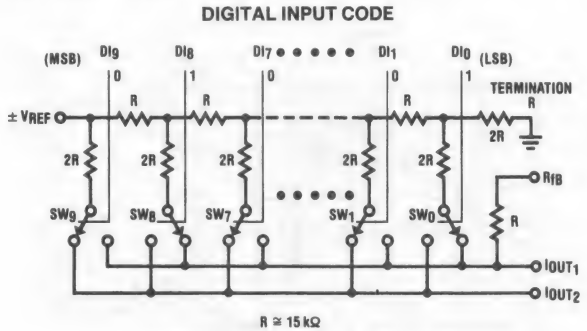
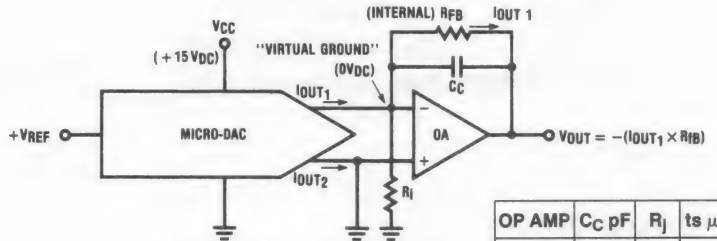


FIGURE 2. Current Mode Switching



OP AMP	C_C pF	R_I	ts μS
LF356	22	∞	3
LF351	24	∞	4
LF357	10	2.4k	1.5

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FIGURE 3. Converting I_{OUT} to V_{OUT}

shown in *Figure 4*, where the sign of the reference voltage has been changed to provide a positive output voltage. Note that the output current, I_{OUT1}, now flows through the R_{FB} pin.

5.1.2 Providing a Bipolar Output Voltage with the DAC in the Current Switching Mode

The addition of a second op amp to the circuit of *Figure 4* can be used to generate a bipolar output voltage from a fixed reference voltage *Figure 5*. This, in effect, gives sign significance to the MSB of the digital input word to allow two quadrant multiplication of the reference voltage. The polarity of the reference can also be reversed to realize the full four-quadrant multiplication.

The applied digital word is offset binary which includes a code to output zero volts without the need of a large valued resistor common to existing bipolar multiplying DAC circuits. Offset binary code can be derived from 2's complement data (most common for signed processor arithmetic) by inverting the state of the MSB in either software or hardware. After doing this the output then responds in accordance to the following expression:

$$V_O = V_{REF} \times \frac{D}{512}$$

where V_{REF} can be positive or negative and D is the signed decimal equivalent of the 2's complement processor data. ($-512 \le D \le +511$ or $1000000000 \le D \le 0111111111$). If the applied digital input is interpreted as the decimal equivalent of a true binary word, V_{OUT} can be found by:

$$V_O = V_{REF}\left(\frac{D-512}{512}\right) \qquad 0 \le D \le 1023$$

With this configuration, only the offset voltage of amplifier 1 need be nulled to preserve linearity of the DAC. The offset voltage error of the second op amp has no effect on linearity. It presents a constant output voltage error and should be nulled only if absolute accuracy is needed. Another advantage of this configuration is that the values of the external resistors required do not have to match the value of the internal DAC resistors; they need only to match and temperature track each other.

A thin film 4 resistor network available from Beckman Instruments, Inc. (part no. 694-3-R10K-D) is ideally suited for this application. Two of the four available 10 kΩ resistor can be paralleled to form R in *Figure 5* and the other two can be used separately as the resistors labeled 2R.

Operation is summarized in the table below:

2's Comp. (Decimal)	2's Comp. (Binary)	Applied Digital Input	Applied True Binary (Decimal)	V_{OUT} $+V_{REF}$	V_{OUT} $-V_{REF}$
+511	0111111111	1111111111	1023	$V_{REF} - 1$ LSB	$-\lvert V_{REF}\rvert + 1$ LSB
+256	0100000000	1100000000	768	$V_{REF}/2$	$-\lvert V_{REF}\rvert/2$
0	0000000000	1000000000	512	0	0
−1	1111111111	0111111111	511	−1 LSB	+1 LSB
−256	1100000000	0100000000	256	$-V_{REF}/2$	$+\lvert V_{REF}\rvert/2$
−512	1000000000	0000000000	0	$-V_{REF}$	$+\lvert V_{REF}\rvert$

with: $1\ LSB = \frac{|V_{REF}|}{512}$

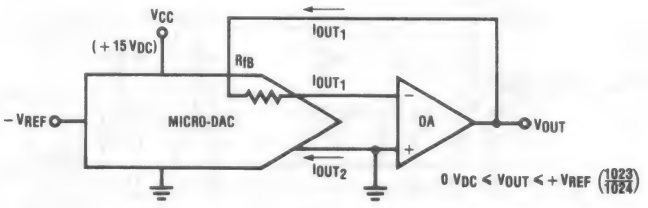

FIGURE 4. Providing a Unipolar Output Voltage

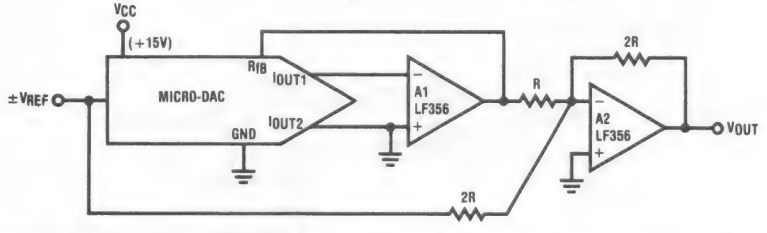

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FIGURE 5. Providing a Bipolar Output Voltage with the DAC in the Current Switching Mode

5.2 Analog Operation in the Voltage Switching Mode

Some useful application circuits result if the R-2R ladder is operated in the voltage switching mode. There are two very important things to remember when using the DAC in the voltage mode. The reference voltage (+V) must always be positive since there are parasitic diodes to ground on the I_{OUT1} pin which would turn on if the reference voltage went negative. To maintain a degradation of linearity less than ±0.005%, keep $+V \leq 3\ V_{DC}$ and V_{CC} at least 10V more positive than +V. *Figures 6* and *7* show these errors for the voltage switching mode. This operation appears unusual, since a reference voltage (+V) is applied to the I_{OUT1} pin and the voltage output is the V_{REF} pin. This basic idea is shown in *Figure 8*.

This V_{OUT} range can be scaled by use of a non-inverting gain stage as shown in *Figure 9*.

Notice that this is unipolar operation since all voltages are positive. A bipolar output voltage can be obtained by using a single op amp as shown in *Figure 10*. For a digital input code of all zeros, the output voltage from the V_{REF} pin is zero volts. The external op amp now has a single input of +V and is operating with a gain of −1 to this input. The output of the op amp therefore will be at −V for a digital input of all zeros. As the digital code increases, the output voltage at the V_{REF} pin increases.

Notice that the gain of the op amp to voltages which are applied to the (+) input is +2 and the gain to voltages which are applied to the input resistor, R, is −1. The output voltage of the op amp depends on both of these inputs and is given by:

$$V_{OUT} = (+V)(-1) + V_{REF}(+2)$$

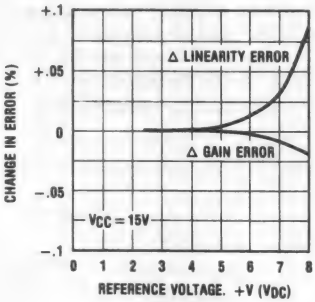

FIGURE 6

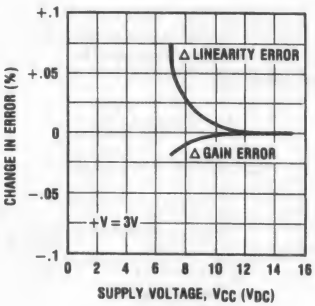

FIGURE 7

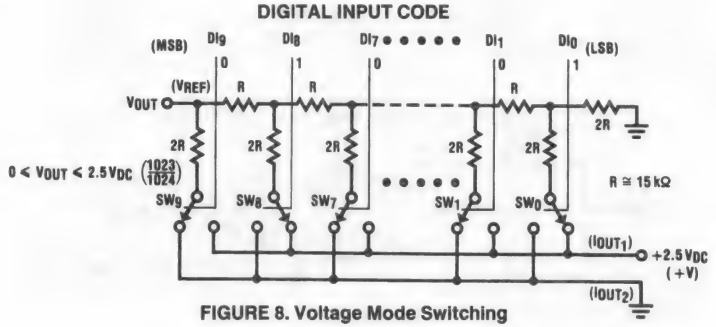

FIGURE 8. Voltage Mode Switching

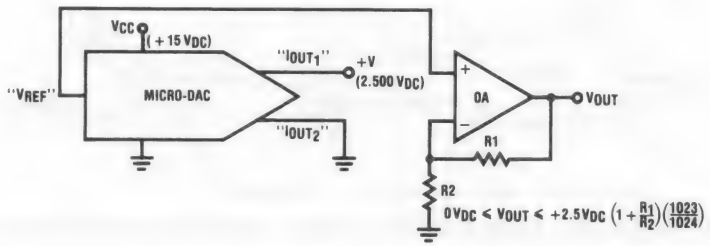

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FIGURE 9. Amplifying the Voltage Mode Output (Single Supply Operation)

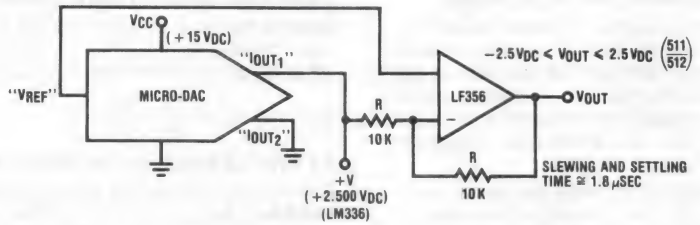
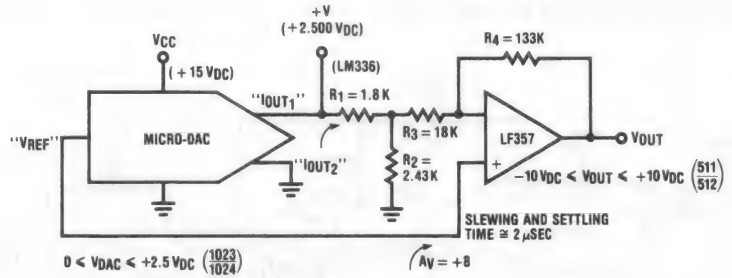


FIGURE 10. Providing a Bipolar Output Voltage with a Single Op Amp



TL/H/5688–13

FIGURE 11. Increasing the Output Voltage Swing

The output voltage swing can be expanded by adding 2 resistors to *Figure 10* as shown in *Figure 11*. These added resistors are used to attenuate the +V voltage. The overall gain, $A_V(-)$, from the +V terminal to the output of the op amp determines the most negative output voltage, −4(+V) (when the V_{REF} voltage at the + input of the op amp is zero) with the component values shown. The complete dynamic range of V_{OUT} is provided by the gain from the (+) input of the op amp. As the voltage at the V_{REF} pin ranges from 0V to +V(1023/1024) the output of the op amp will range from −10 V_{DC} to +10V (1023/1024) when using a +V voltage of +2.500 V_{DC}. The 2.5 V_{DC} reference voltage can be easily developed by using the LM336 zener which can be biased through the R_{FB} internal resistor, connected to V_{CC}.

5.3 Op Amp V_{OS} Adjust (Zero Adjust) for Current Switching Mode

Proper operation of the ladder requires that all of the 2R legs always go to exactly 0 V_{DC} (ground). Therefore offset voltage, V_{OS}, of the external op amp cannot be tolerated as every millivolt of V_{OS} will introduce 0.01% of added linearity error. At first this seems unusually sensitive, until it becomes clear the 1 mV is 0.01% of the 10V reference! High resolution converters of high accuracy require attention to every detail in an application to achieve the available performance which is inherent in the part. To prevent this source of error, the V_{OS} of the op amp has to be initially zeroed. This is the "zero adjust" of the DAC calibration sequence and should be done first.

If the V_{OS} is to be adjusted there are a few points to consider. Note that no "dc balancing" resistance should be used in the grounded positive input lead of the op amp. This resistance and the input current of the op amp can also create errors. The low input biasing current of the BI-FET op amps makes them ideal for use in DAC current to voltage applications. The V_{OS} of the op amp should be adjusted with a digital input of all zeros to force I_{OUT} = 0 mA. A 1 kΩ resistor can be temporarily connected from the inverting input to ground to provide a dc gain of approximately 15 to the V_{OS} of the op amp and make the zeroing easier to sense.

5.4 Full-Scale Adjust

The full-scale adjust procedure depends on the application circuit and whether the DAC is operated in the current switching mode or in the voltage switching mode. Techniques are given below for all of the possible application circuits.

5.4.1 Current Switching with Unipolar Output Voltage

After doing a "zero adjust," set all of the digital input levels HIGH and adjust the magnitude of V_{REF} for

$$V_{OUT} = -(\text{ideal } V_{REF}) \frac{1023}{1024}$$

This completes the DAC calibration.

5.4.2 Current Switching with Bipolar Output Voltage

The circuit of *Figure 12* shows the 3 adjustments needed. The first step is to set all of the digital inputs LOW (to force I_{OUT1} to 0) and then trim "zero adj." for zero volts at the inverting input (pin 2) of 0A1. Next, with a code of all zeros still applied, adjust "−FS adj.", the reference voltage, for $V_{OUT} = \pm|(\text{ideal } V_{REF})|$. The sign of the output voltage will be opposite that of the applied reference.

Finally, set all of the digital inputs HIGH and adjust "+FS adj." for $V_{OUT} = V_{REF}$ (511/512). The sign of the output at this time will be the same as that of the reference voltage. The addition of the 200Ω resistor in series with the V_{REF} pin of the DAC is to force the circuit gain error from the DAC to be negative. This insures that adding resistance to R_{fb}, with the 500Ω pot, will always compensate the gain error of the DAC.

5.4.3 Voltage Switching with a Unipolar Output Voltage

Refer to the circuit of *Figure 13* and set all digital inputs LOW. Trim the "zero adj." for $V_{OUT} = 0\ V_{DC} \pm 1$ mV. Then set all digital inputs HIGH and trim the "FS Adj." for:

$$V_{OUT} = (+V)\left(1 + \frac{R_1}{R_2}\right)\frac{1023}{1024}$$

5.4.4 Voltage Switching with a Bipolar Output Voltage

Refer to *Figure 14* and set all digital inputs LOW. Trim the "−FS Adj." for $V_{OUT} = -2.5\ V_{DC}$. Then set all digital inputs HIGH and trim the "+FS Adj." for $V_{OUT} = +2.5$ (511/512) V_{DC}. Test the zero by setting the MS digital input HIGH and all the rest LOW. Adjust V_{OS} of amp #3, if necessary, and recheck the full-scale values.

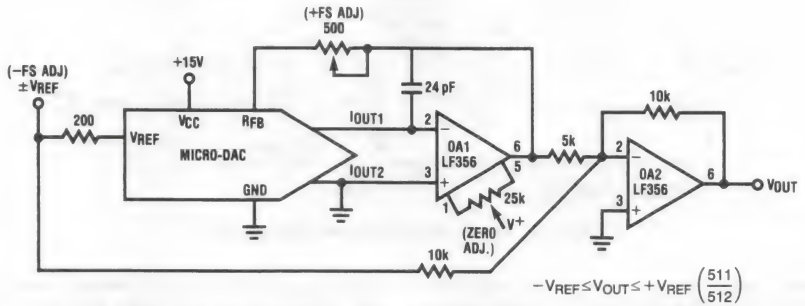
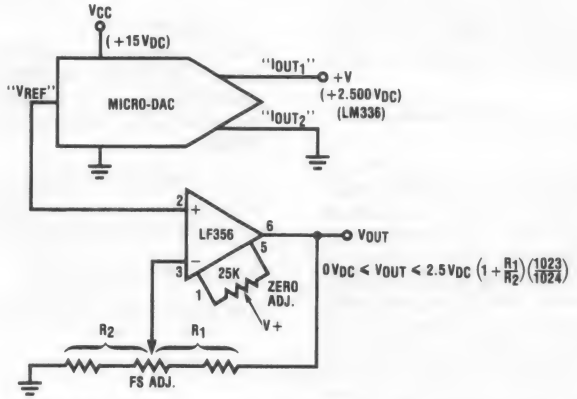
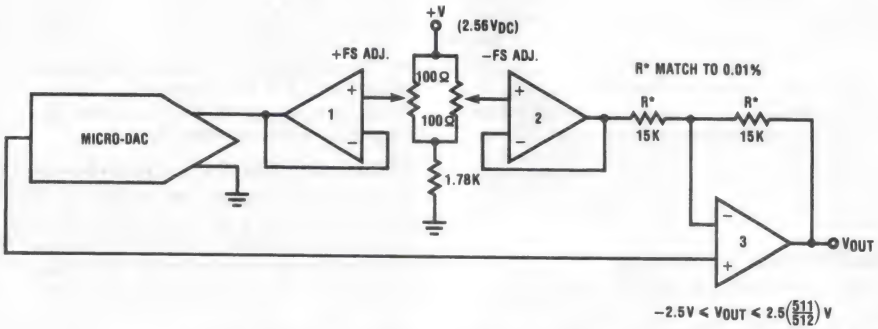


FIGURE 12. Full Scale Adjust — Current Switching with Bipolar Output Voltage



TL/H/5688–14

FIGURE 13. Full Scale Adjust — Voltage Switching with a Unipolar Output Voltage



TL/H/5688-15

FIGURE 14. Voltage Switching with a Bipolar Output Voltage

6.0 DIGITAL CONTROL DESCRIPTION

The DAC1000 series of products can be used in a wide variety of operating modes. Most of the options are shown in Table 1. Also shown in this table are the section numbers of this data sheet where each of the operating modes is discussed. For example, if your main interest in interfacing to a μP with an 8-bit data bus you will be directed to Section 6.1.0.

The first consideration is "will the DAC be interfaced to a μP with an 8-bit or a 16-bit data bus or used in the stand-alone mode?" For the 8-bit data bus, a second selection is made on how the 2nd digital data buffer (the DAC Latch) is updated by a transfer from the 1st digital data buffer (the Input Latch). Three options are provided: 1) an automatic transfer when the 2nd data byte is written to the DAC, 2) a transfer which is under the control of the μP and can include more than one DAC in a simultaneous transfer, or 3) a transfer which is under the control of external logic. Further, the data format can be either left justified or right justified.

When interfacing to a μP with a 16-bit data bus only two selections are available: 1) operating the DAC with a single digital data buffer (the transfer of one DAC does not have to be synchronized with any other DACs in the system), or 2) operating with a double digital data buffer for simultaneous transfer, or updating, of more than one DAC.

For operating without a μP in the stand alone mode, three options are provided: 1) using only a single digital data buffer, 2) using both digital data buffers — "double buffered," or 3) allowing the input digital data to "flow through" to provide the analog output without the use of any data latches.

To reduce the required reading, only the applicable sections of 6.1 through 6.4 need be considered.

6.1 Interfacing to an 8-Bit Data Bus

Transferring 10 bits of data over an 8-bit bus requires two write cycles and provides four possible combinations which depend upon two basic data format and protocol decisions:

1. Is the data to be left justified (considered as fractional binary data with the binary point to the left) or right justified (considered as binary weighted data with the binary point to the right)?
2. Which byte will be transferred first, the most significant byte (MS byte) or the least significant byte (LS byte)?

Table 1

Operating Mode	Automatic Transfer			μP Control Transfer			External Transfer		
Data Bus	Section	Figure No. (24-Pin)	Figure No. (20-Pin)	Section	Figure No. (24-Pin)	Figure No. (20-Pin)	Section	Figure No. (24-Pin)	Figure No. (20-Pin)
8-Bit Data Bus (6.1.0)									
Right Justified (6.1.1)	6.2.1	16		6.2.2	16		6.2.3	16	
Left Justified (6.1.2)	6.2.1	17	18	6.2.2	17	18	6.2.3	17	18
16-Bit Data Bus (6.3.0)	Single Buffered			Double Buffered			Flow Through		
	6.3.1	19	20	6.3.2	19	20	Not Applicable		
Stand Alone (6.4.0)	Single Buffered			Double Buffered			Flow Through		
	6.4.1	19	20	6.4.2	19	20	6.4.3	19	NA

These data possibilities are shown in *Figure 15*. Note that the justification of data depends on how the 10-bit data word is located within the 16-bit data source (CPU) register. In either case, there is a surplus of 6 bits and these are shown as "don't care" terms ("×") in this figure.

All of these DACs load 10 bits on the 1st write cycle. A particular set of 2 bits is then overwritten on the 2nd write cycle, depending on the justification of the data. This requires the 1st write cycle to contain the LS or LO Byte data group for all right justified data options. For all left justified data options, the 1st write cycle must contain the MS or Hi Byte data group.

6.1.1 Providing for Optional Data Format

The DAC1000/1/2 (24-pin parts) can be used for either data formatting by tying the LJ/$\overline{RJ}$ pin either high or low, respectively. A simplified logic diagram which shows the external connections to the data bus and the internal functions of both of the data buffer registers (Input Latch and DAC Register) is shown in *Figure 16* for the right justified data operation. *Figure 17* is for left justified data.

6.1.2 For Left Justified Data

For applications which require left justified data, DAC1006–1008 (20-pin parts) can be used. A simplified logic diagram which shows the external connections to the data bus and the internal functions of both of the data buffer registers (Input Latch and DAC Register) is shown in *Figure 18*. These parts require the MS or Hi Byte data group to be transferred on the 1st write cycle.

6.2 Controlling Data Transfer for an 8-Bit Data Bus

Three operating modes are possible for controlling the transfer of data from the Input Latch to the DAC Register, where it will update the analog output voltage. The simplest is the automatic transfer mode, which causes the data transfer to occur at the time of the 2nd write cycle. This is recommended when the exact timing of the changes of the DAC analog output are not critical. This typically happens where each DAC is operating individually in a system and the analog updating of one DAC is not required to be synchronized to any other DAC. For synchronized DAC updating, two options are provided: μP control via a common $\overline{XFER}$ strobe or external update timing control via an external strobe. The details of these options are now shown.

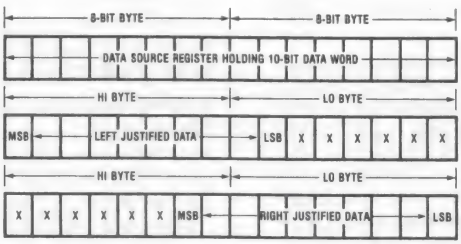


FIGURE 15. Fitting a 10-Bit Data Word into 16 Available Bit Locations

DAC1000/1001/1002 (24-Pin Parts)

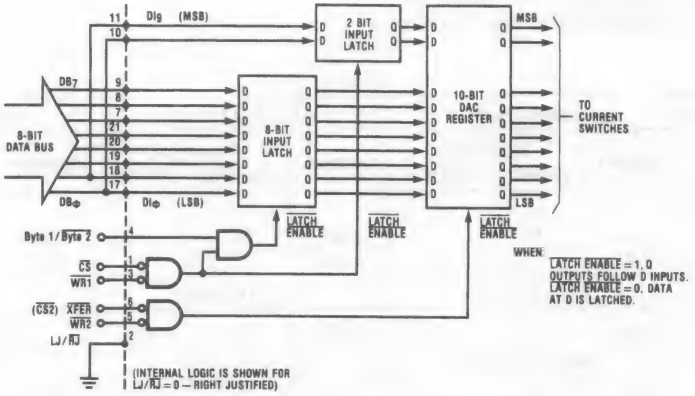


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FIGURE 16. Input Connections and Controls for DAC1000–1002 Right Justified Data Option

DAC1000/1001/1002 (24-Pin Parts)

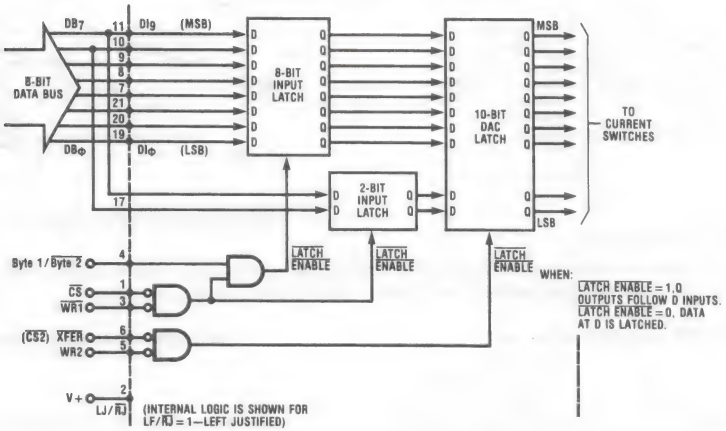
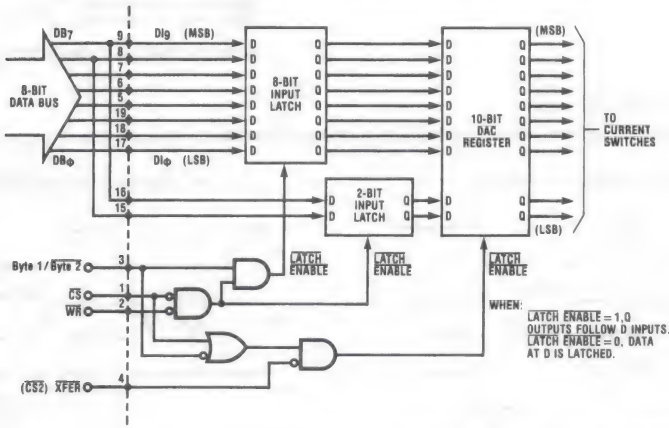


FIGURE 17. Input Connections and Controls for DAC1000–1002 Left Justified Data Option

DAC1006/1007/1008 (20-Pin Parts for Left Justified Data)



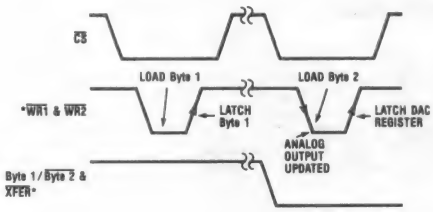
TL/H/5688–17

FIGURE 18. Input Connections and Controls for DAC1006/1007/1008 Left Justified Data

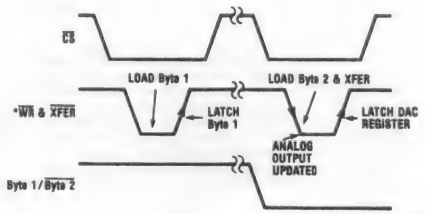
6.2.1 Automatic Transfer

This makes use of a double byte (double precision) write. The first byte (8 bits) is strobed into the input latch and the second byte causes a simultaneous strobe of the two remaining bits into the input latch and also the transfer of the complete 10-bit word from the input latch to the DAC register. This is shown in the following timing diagrams; the point in time where the analog output is updated is also indicated on these diagrams.

DAC1000/1001/1002 (24-Pin Parts)



DAC1006/1007/1008 (20-Pin Parts)



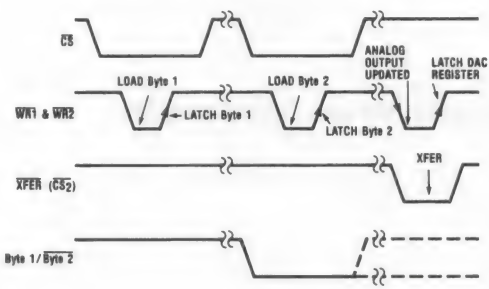
TL/H/5688–18

*SIGNIFIES CONTROL INPUTS WHICH ARE DRIVEN IN PARALLEL

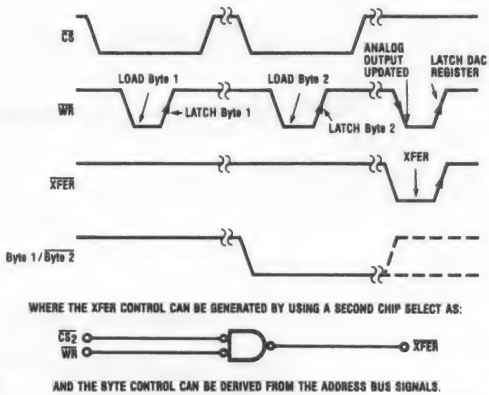
6.2.2 Transfer Using μP Write Stroke

The input latch is loaded with the first two write strobes. The $\overline{XFER}$ signal is provided by external logic, as shown below, to cause the transfer to be accomplished on a third write strobe. This is shown in the following diagrams:

DAC1000/1001/1002 (24-Pin Parts)



DAC1006/1007/1008 (20-Pin Parts)



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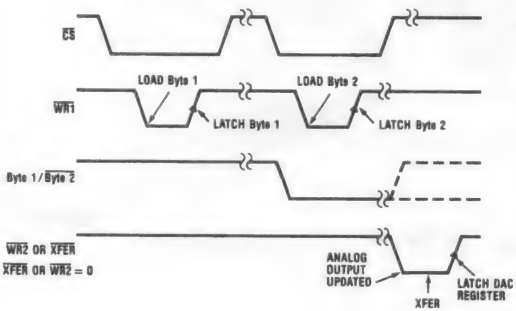
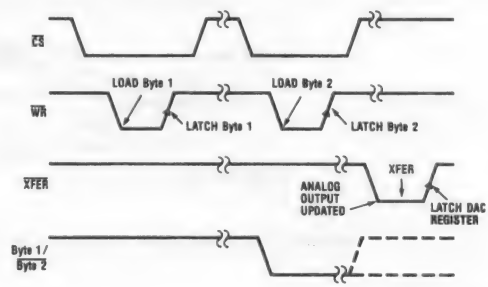
6.2.3 Transfer Using an External Strobe

This is similar to the previous operation except the $\overline{XFER}$ signal is not provided by the μP. The timing diagram for this is:

DAC1000/1001/1002 (24-Pin Parts)



DAC1006/1007/1008 (20-Pin Parts)



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6.3 Interfacing to a 16-Bit Data Bus

The interface to a 16-bit data bus is easily handled by connecting to 10 of the available bus lines. This allows a wiring selected right justified or left justified data format. This is shown in the connection diagrams of *Figures 19* and *20*, where the use of DB6 to DB15 gives left justified data operation. Note that any part number can be used and the Byte1/$\overline{\text{Byte2}}$ control should be wired Hi.

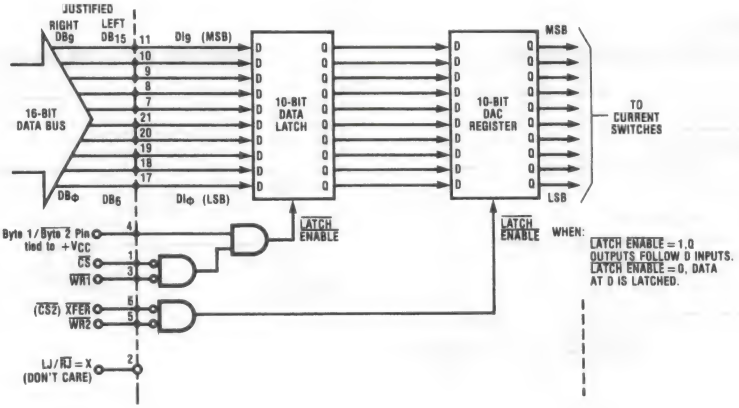


FIGURE 19. Input Connections and Logic for DAC1000–1002 with 16-Bit Data Bus

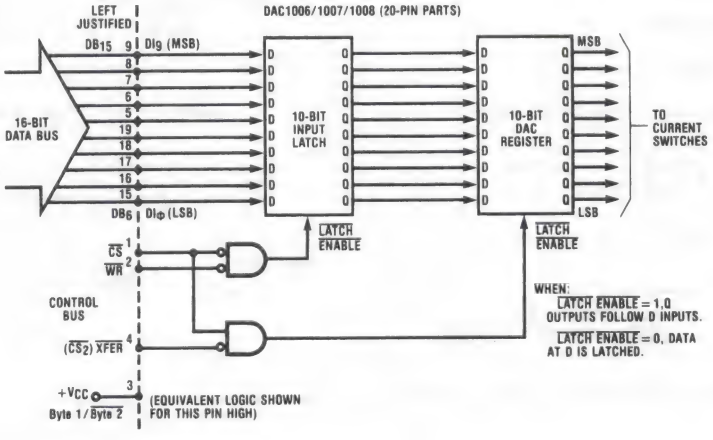
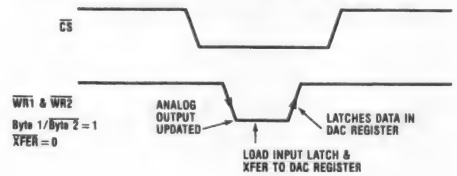


FIGURE 20. Input Connections and Logic for DAC1006/1007/1008 with 16-Bit Data Bus

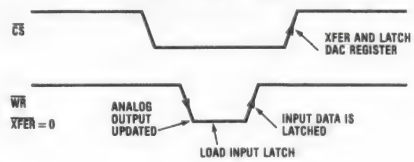
Three operating modes are possible: flow through, single buffered, or double buffered. The timing diagrams for these are shown below:

6.3.1 Single Buffered

DAC1000/1001/1002 (24-Pin Parts)

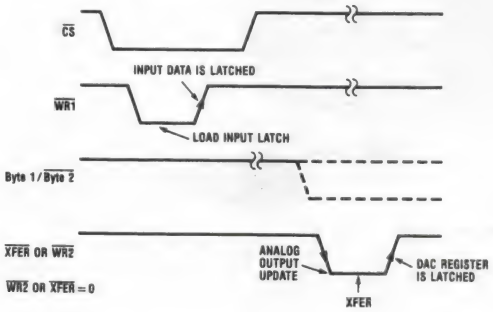


DAC1006/1007/1008 (20-Pin Parts)

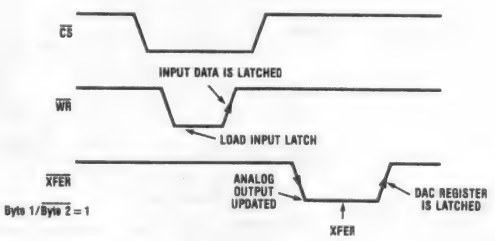


6.3.2 Double Buffered

DAC 1000/1001/1002 (24-Pin Parts)



DAC1006/1007/1008 (20-Pin Parts)



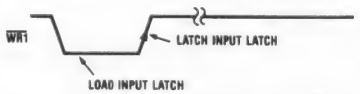
TL/H/5688–22

6.4 Stand Alone Operation

For applications for a DAC which are not under μP control (stand alone) there are two basic operating modes, single buffered and double buffered. The timing diagrams for these are shown below:

6.4.1 Single Buffered

DAC1000/1001/1002 (24-Pin Parts)

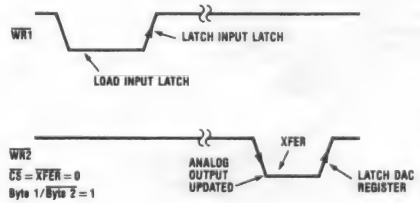


DAC1006/1007/1008 (20-Pin Parts)

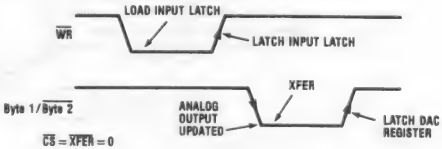
XFER TO DAC REGISTER
Byte 1/Byte 2
LOAD INPUT LATCH
ANALOG OUTPUT/ UPDATED
LATCHES DATA IN DAC REGISTER (INPUT DATA MUST REMAIN VALID UNTIL THIS TIME)

6.4.2 Double Buffered

DAC1000/1001/1002 (24-Pin Parts)



DAC1006/1007/1008 (20-Pin Parts)*



TL/H/5688–23

*For a connection diagram of this operating mode use *Figure 18* for the Logic and *Figure 20* for the Data Input connections.

6.4.3 Flow Through

This operating mode causes the 10-bit input word to directly create the DAC output without any latching involved.

DAC1000/1001/1002 (24-Pin Parts)

$\overline{WR1} = \overline{WR2} = \overline{CS} = \overline{XFER} = 0$

Byte 1/$\overline{Byte\ 2}$ = 1

7.0 MICROPROCESSOR INTERFACE

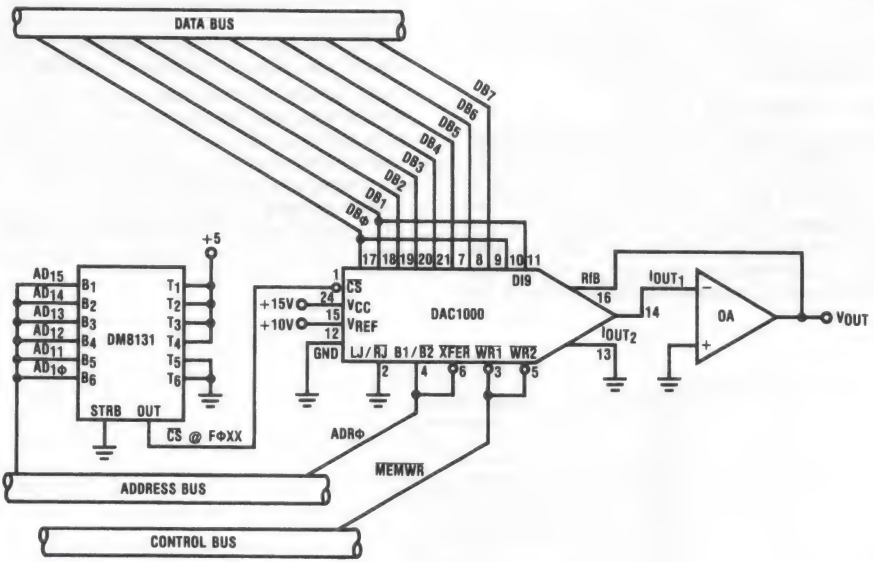
The logic functions of the DAC1000 family have been oriented towards an ease of interface with all popular μPs. The following sections discuss in detail a few useful interface schemes.

7.1 DAC1001/1/2 to INS8080A Interface

Figure 21 illustrates the simplicity of interfacing the DAC1000 to an INS8080A based microprocessor system.

The circuit will perform an automatic transfer of the 10 bits of output data from the CPU to the DAC register as outlined in Section 6.2.1, "Controlling Data Transfer for an 8-Bit Data Bus."

Since a double byte write is necessary to control the DAC with the INS8080A, a possible instruction to achieve this is a PUSH of a register pair onto a "stack" in memory. The 16-bit register pair word will contain the 10 bits of the eventual DAC input data in the proper sequence to conform to both



TL/H/5688–24

NOTE: DOUBLE BYTE STORES CAN BE USED.
e.g. THE INSTRUCTION SHLD F001 STORES THE L REG INTO B1 AND THE H REG INTO B2 AND TRANSFERS THE RESULT TO THE DAC REGISTER. THE OPERAND OF THE SHLD INSTRUCTION MUST BE AN ODD ADDRESS FOR PROPER TRANSFER.

FIGURE 21. Interfacing the DAC1000 to the INS8080A CPU Group

the requirements of the DAC (with regard to right or left justified data) and the implementation of the PUSH instruction which will output the higher order byte of the register pair (i.e., register B of the BC pair) first. The DAC will actually appear as a two-byte "stack" in memory to the CPU. The auto-decrementing of the stack pointer during a PUSH allows using address bit 0 of the stack pointer as the Byte1/$\overline{\text{Byte2}}$ and $\overline{\text{XFER}}$ strobes if bit 0 of the stack pointer address −1, (SP−1), is a "1" as presented to the DAC. Additional address decoding by the DM8131 will generate a unique DAC chip select (CS) and synchronize this CS to the two memory write strobes of the PUSH instruction.

To reset the stack pointer so new data may be output to the same DAC, a POP instruction followed by instructions to insure that proper data is in the DAC data register pair before it is "PUSHED" to the DAC should be executed, as the POP instruction will arbitrarily alter the contents of a register pair.

Another double byte write instruction is Store H and L Direct (SHLD), where the HL register pair would temporarily contain the DAC data and the two sequential addresses for the DAC are specified by the instruction op code. The auto incrementing of the DAC address by the SHLD instruction permits the same simple scheme of using address bit 0 to generate the byte number and transfer strobes.

7.2 DAC1000 to MC6820/1 PIA Interface

In *Figure 22* the DAC1000 is interfaced to an M6800 system through an MC6820/1 Peripheral Interface Adapter (PIA). In this case the CS pin of the DAC is grounded since the PIA is already mapped in the 6800 system memory space and no decoding is necessary. Furthermore, by using both Ports A and B of the PIA the 10-bit data transfer, assumed right justified again in two 8-bit bytes, is greatly simplified. The HIGH byte is loaded into Output Register A (ORA) of the PIA, and the LOW byte is loaded into ORB. The 10-bit data transfer to the DAC and the corresponding analog output change occur simultaneously upon CB2 going LOW under program control. The 10-bit data word in the DAC register will be latched (and hence V_{OUT} will be fixed) when CB2 is brought back HIGH.

If both output ports of the PIA are not available, it is possible to interface the DAC1000 through a single port without much effort. However, additional logic at the CB2(or CA2) lines or access to some of the 6800 system control lines will be required.

7.3 Noise Considerations

A typical digital/microprocessor bus environment is a tremendous potential source of high frequency noise which can be coupled to sensitive analog circuitry. The fast edges of the data and address bus signals generate frequency components of 10's of megahertz and can cause noise spikes to appear at the DAC output. These noise spikes occur when the data bus changes state or when data is transferred between the latches of the device.

In low frequency or DC applications, low pass filtering can reduce these noise spikes. This is accomplished by overcompensating the DAC output amplifier by increasing the value of the feedback capacitor (C_C in *Figure 3*).

In applications requiring a fast transient response from the DAC and op amp, filtering may not be feasible. Adding a latch, DM74LS374, as shown in *Figure 23* isolates the device from the data bus, thus eliminating noise spikes that occur every time the data bus changes state. Another method for eliminating noise spikes is to add a sample and hold after the DAC op amp. This also has the advantage of eliminating noise spikes when changing digital codes.

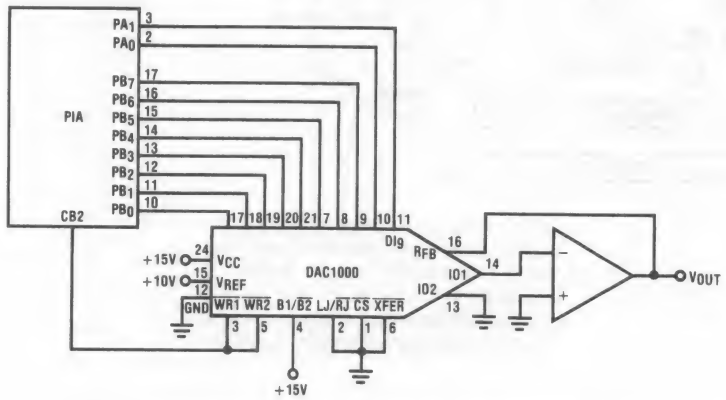


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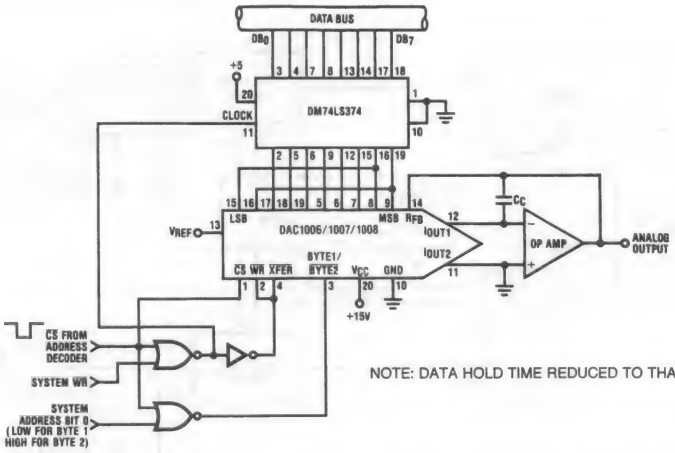
FIGURE 22. DAC1000 to MC6820/1 PIA Interface



NOTE: DATA HOLD TIME REDUCED TO THAT OF DM74LS374 (≈10 ns)

FIGURE 23. Isolating Data Bus from DAC Circuitry to Eliminate Digital Noise Coupling

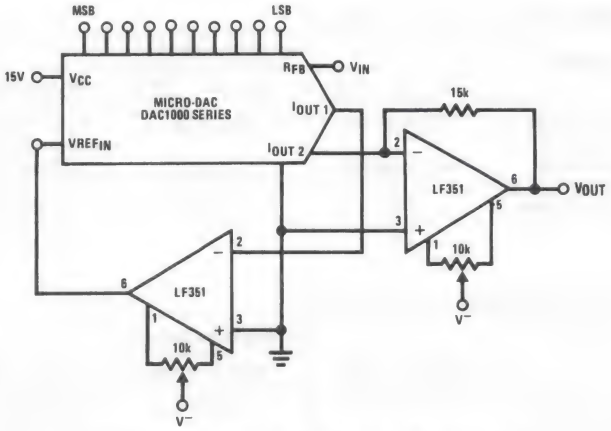


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FIGURE 24. Digitally Controlled Amplifier/Attenuator

7.4 Digitally Controlled Amplifier/Attenuator

An unusual application of the DAC, *Figure 24*, applies the input voltage via the on-chip feedback resistor. The lower op amp automatically adjusts the $V_{REF\ IN}$ voltage such that I_{OUT1} is equal to the input current (V_{IN}/Rf_B). The magnitude of this $V_{REF\ IN}$ voltage depends on the digital word which is in the DAC register. I_{OUT2} then depends upon both the magnitude of V_{IN} and the digital word. The second op amp converts I_{OUT2} to a voltage, V_{OUT}, which is given by:

$$V_{OUT} = V_{IN}\left(\frac{1023-N}{N}\right), \text{ where } 0<N\leq 1023.$$

Note that N = 0 (or a digital code of all zeros) is not allowed or this will cause the output amplifier to saturate at either $\pm V_{MAX}$, depending on the sign of V_{IN}.

To provide a digitally controlled divider, the output op amp can be eliminated. Ground the I_{OUT2} pin of the DAC and V_{OUT} is now taken from the lower op amp (which also drives the V_{REF} input of the DAC). The expression for V_{OUT} is now given by

$$V_{OUT} = -\frac{V_{IN}}{M} \text{ where M = Digital input (expressed as a fractional binary number).}$$

$$0<M<1.$$

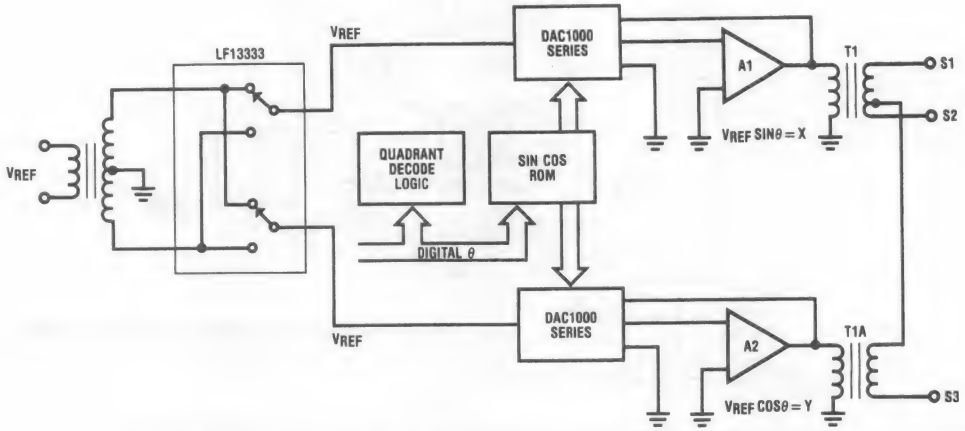

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FIGURE 25. Digital to Synchro Converter

Ordering Information

1. All Logic Features — 24-pin package.

Accuracy	Temperature Range		
	−40°C to +85°C	−55°C to +125°C	0° to +70°C
0.05% (10-bit)	DAC1000LCJ	DAC1000LJ	DAC 1000LCN
0.10% (9-bit)			DAC1001LCN
0.20% (8-bit)	DAC1002LCJ	DAC1002LJ	DAC1002LCN
Package Outline	J24A	J24A	N24A

2. For Left Justified Data — 20-pin package.

Accuracy	Temperature Range		
	−40°C to +85°C	−55°C to +125°C	0° to +70°C
0.05% (10-bit)	DAC1006LCJ	DAC1006LJ	DAC1006LCN
0.10% (9-bit)			DAC1007LCN
0.20% (8-bit)	DAC1008LCJ	DAC1008LJ	DAC1008LCN
Package Outline	J20A	J20A	N20A

National Semiconductor

DAC1020/DAC1021/DAC1022
10-Bit Binary Multiplying D/A Converter
DAC1220/DAC1221/DAC1222
12-Bit Binary Multiplying D/A Converter

General Description

The DAC1020 and the DAC1220 are, respectively, 10 and 12-bit binary multiplying digital-to-analog converters. A deposited thin film R-2R resistor ladder divides the reference current and provides the circuit with excellent temperature tracking characteristics (0.0002%/°C linearity error temperature coefficient maximum). The circuit uses CMOS current switches and drive circuitry to achieve low power consumption (30 mW max) and low output leakages (200 nA max). The digital inputs are compatible with DTL/TTL logic levels as well as full CMOS logic level swings. This part, combined with an external amplifier and voltage reference, can be used as a standard D/A converter; however, it is also very attractive for multiplying applications (such as digitally controlled gain blocks) since its linearity error is essentially independent of the voltage reference. All inputs are protected from damage due to static discharge by diode clamps to V^+ and ground.

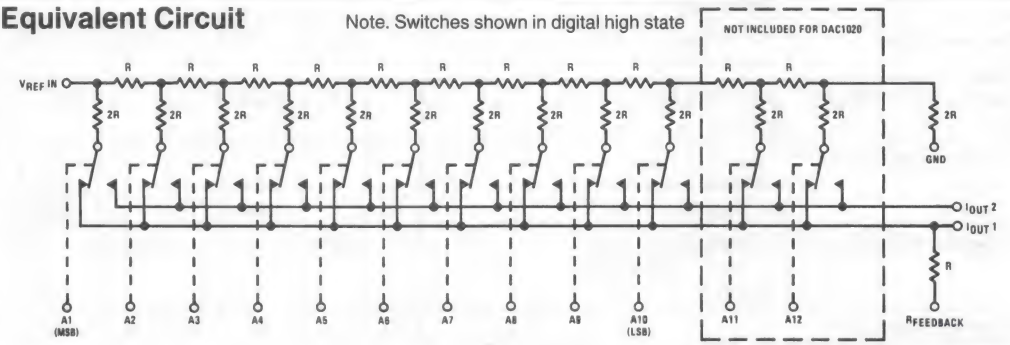
This part is available with 10-bit (0.05%), 9-bit (0.10%), and 8-bit (0.20%) non-linearity guaranteed over temperature (note 1 of electrical characteristics). The DAC1020, DAC1021 and DAC1022 are direct replacements for the 10-bit resolution AD7520 and AD7530 and equivalent to the AD7533 family. The DAC1220, DAC1221 and DAC1222 are direct replacements for the 12-bit resolution AD7521 and AD7531 family.

Features

- Linearity specified with zero and full-scale adjust only
- Non-linearity guaranteed over temperature
- Integrated thin film on CMOS structure
- 10-bit or 12-bit resolution
- Low power dissipation 10 mW @15V typ
- Accepts variable or fixed reference $-25V \leq V_{REF} \leq 25V$
- 4-quadrant multiplying capability
- Interfaces directly with DTL, TTL and CMOS
- Fast settling time—500 ns typ
- Low feedthrough error—½ LSB @100 kHz typ

Equivalent Circuit

Note. Switches shown in digital high state



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Ordering Information

10-BIT D/A CONVERTERS

Temperature Range		0°C to 70°C		−40°C to +85°C		−55°C to +125°C	
Non-Linearity	0.05%	DAC1020LCN	AD7520LN,AD7530LN	DAC1020LCJ	AD7520LD,AD7530LD	DAC1020LJ	AD7520UD
	0.10%	DAC1021LCN	AD7520KN,AD7530KN	DAC1021LCJ	AD7520KD,AD7530KD	DAC1021LJ	AD7520TD
	0.20%	DAC1022LCN	AD7520JN,AD7530JN	DAC1022LCJ	AD7520JD,AD7530JD	DAC1022LJ	AD7520SD
Package Outline		N16A		J16A		J16A	

12-BIT D/A CONVERTERS

Temperature Range		0°C to 70°C		−40°C to +85°C		−55°C to +125°C	
Non-Linearity	0.05%	DAC1220LCN	AD7521LN,AD7531LN	DAC1220LCJ	AD7521LD,AD7531LD	DAC1220LJ	AD7521UD
	0.10%	DAC1221LCN	AD7521KN,AD7531KN				
	0.20%	DAC1222LCN	AD7521JN,AD7531JN	DAC1222LCJ	AD7521JD,AD7531JD	DAC1222LJ	AD7521SD
Package Outline		N18A		J18A		J18A	

Note. Devices may be ordered by either part number.

4

Absolute Maximum Ratings (Note 5)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

V^+ to Gnd	17V
V_{REF} to Gnd	±25V
Digital Input Voltage Range	V^+ to Gnd
DC Voltage at Pin 1 or Pin 2 (Note 3)	−100 mV to V^+
Storage Temperature Range	−65°C to +150°C
Lead Temperature (Soldering, 10 sec.)	
Dual-In-Line Package (plastic)	260°C
Dual-In-Line Package (ceramic)	300°C
ESD Susceptibility (Note 4)	800V

Operating Ratings

	Min	Max	Units
Temperature (T_A)			
DAC1020LJ, DAC1021LJ	−55	+125	°C
DAC1022LJ, DAC1220LJ	−55	+125	°C
DAC1222LJ	−55	+125	°C
DAC1020LCJ, DAC1021LCJ	−40	+85	°C
DAC1022LCJ, DAC1220LCJ	−40	+85	°C
DAC1222LCJ	−40	+85	°C
DAC1020LCN, DAC1021LCN	0	+70	°C
DAC1022LCN, DAC1220LCN	0	+70	°C
DAC1221LCN, DAC1222LCN	0	+70	°C

Electrical Characteristics ($V^+ = 15V$, $V_{REF} = 10.000V$, $T_A = 25°C$ unless otherwise specified)

Parameter	Conditions	DAC1020, DAC1021, DAC1022			DAC1220, DAC1221, DAC1222			Units
		Min	Typ	Max	Min	Typ	Max	
Resolution		10			12			Bits
Linearity Error	$T_{MIN} < T_A < T_{MAX}$, $-10V < V_{REF} < +10V$, (Note 1) End Point Adjustment Only (See Linearity Error in Definition of Terms)							
10-Bit Parts	DAC1020, DAC1220			0.05			0.05	% FSR
9-Bit Parts	DAC1021, DAC1221			0.10			0.10	% FSR
8-Bit Parts	DAC1022, DAC1222			0.20			0.20	% FSR
Linearity Error Tempco	$-10V \le V_{REF} \le +10V$, (Notes 1 and 2)			0.0002			0.0002	% FS/°C
Full-Scale Error	$-10V \le V_{REF} \le +10V$, (Notes 1 and 2)		0.3	1.0		0.3	1.0	% FS
Full-Scale Error Tempco	$T_{MIN} < T_A < T_{MAX}$, (Note 2)			0.001			0.001	% FS/°C
Output Leakage Current	$T_{MIN} \le T_A \le T_{MAX}$							
$I_{OUT\,1}$	All Digital Inputs Low			200			200	nA
$I_{OUT\,2}$	All Digital Inputs High			200			200	nA
Power Supply Sensitivity	All Digital Inputs High, $14V \le V^+ \le 16V$, (Note 2), *(Figure 2)*		0.005			0.005		% FS/V
V_{REF} Input Resistance		10	15	20	10	15	20	kΩ
Full-Scale Current Settling Time	$R_L = 100\Omega$ from 0 to 99. 95% FS							
	All Digital Inputs Switched Simultaneously		500			500		ns
V_{REF} Feedthrough	All Digital Inputs Low, $V_{REF} = 20$ Vp-p @ 100 kHz			10			10	mVp-p
	J Package (Note 4)		6	9		6	9	mVp-p
	N Package		2	5		2	5	mVp-p
Output Capacitance								
$I_{OUT\,1}$	All Digital Inputs Low		40			40		pF
	All Digital Inputs High		200			200		pF
$I_{OUT\,2}$	All Digital Inputs Low		200			200		pF
	All Digital Inputs High		40			40		pF

Electrical Characteristics ($V^+ = 15V$, $V_{REF} = 10.000V$, $T_A = 25°C$ unless otherwise specified) (Continued)

Parameter	Conditions	DAC1020, DAC1021, DAC1022			DAC1220, DAC1221, DAC1222			Units
		Min	Typ	Max	Min	Typ	Max	
Digital Input	*(Figure 1)*							
Low Threshold	$T_{MIN} < T_A < T_{MAX}$			0.8			0.8	V
High Threshold	$T_{MIN} < T_A < T_{MAX}$	2.4			2.4			V
Digital Input Current	$T_{MIN} \leq T_A \leq T_{MAX}$							
	Digital Input High		1	100		1	100	μA
	Digital Input Low		−50	−200		−50	−200	μA
Supply Current	All Digital Inputs High		0.2	1.6		0.2	1.6	mA
	All Digital Inputs Low		0.6	2		0.6	2	mA
Operating Power Supply Range	*(Figures 1 and 2)*	5		15	5		15	V

Note 1: $V_{REF} = \pm 10V$ and $V_{REF} = \pm 1V$. A linearity error temperature coefficient of 0.0002% FS for a 45°C rise only guarantees 0.009% maximum change in linearity error. For instance, if the linearity error at 25°C is 0.045% FS it could increase to 0.054% at 70°C and the DAC will be no longer a 10-bit part. Note, however, that the linearity error is specified over the device full temperature range which is a more stringent specification since *it includes* the linearity error temperature coefficient.

Note 2: Using internal feedback resistor as shown in *Figure 3*.

Note 3: Both $I_{OUT\,1}$ and $I_{OUT\,2}$ must go to ground or the virtual ground of an operational amplifier. If $V_{REF} = 10V$, every millivolt offset between $I_{OUT\,1}$ or $I_{OUT\,2}$, 0.005% linearity error will be introduced.

Note 4: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 5: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 6: The maximum power dissipation must be derated at elevated temperatures and is dictated by T_{JMAX}, θ_{JA}, and the ambient temepature, T_A. The maximum allowable power dissipation at any temperature is $P_D = (T_{JMAX} - T_A)/\theta_{JA}$ or the number given in the Absolute Maximum Ratings, whichever is lower. For this device, $T_{JMAX} = 125°C$, and the typical junction-to-ambient thermal resistance of the J18 package when board mounted is 85°C/W. For the J16 package, this number increases to 90°C/W, for the N18 package, θ_{JA} is 120°C/W, and for the N16 this number is 125°C/W.

Typical Performance Characteristics

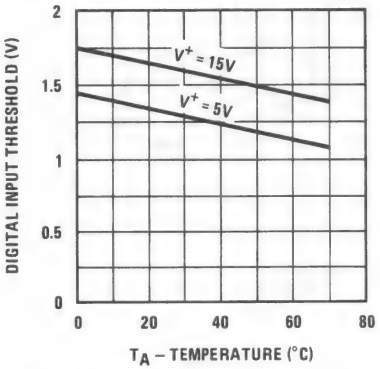

FIGURE 1. Digital Input Threshold vs Ambient Temperature

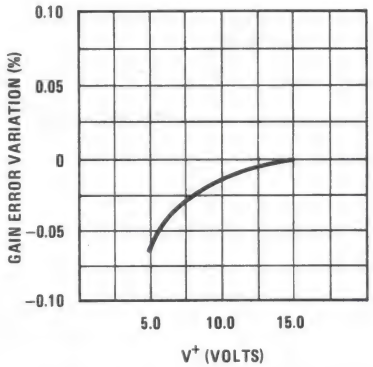

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FIGURE 2. Gain Error Variation vs V^+

Typical Applications

The following applications are also valid for 12-bit systems using the DAC1220 and 2 additional digital inputs.

Operational Amplifier Bias Current (*Figure 3*)

The op amp bias current, I_b, flows through the 15k internal feedback resistor. BI-FET op amps have low I_b and, therefore, the 15k $\times$ I_b error they introduce is negligible; they are strongly recommended for the DAC1020 applications.

V_{OS} Considerations

The output impedance, R_{OUT}, of the DAC is modulated by the digital input code which causes a modulation of the operational amplifier output offset. It is therefore recommended to adjust the op amp V_{OS}. R_{OUT} is $\sim$15k if more than 4 digital inputs are high; R_{OUT} is $\sim$45k if a single digital input is high, and R_{OUT} approaches infinity if all inputs are low.

Operational Amplifier V_{OS} Adjust (*Figure 3*)

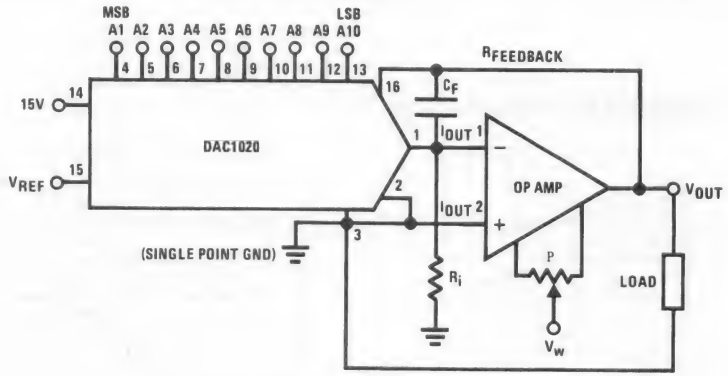
Connect all digital inputs, A1–A10, to ground and adjust the potentiometer to bring the op amp V_{OUT} pin to within ± 1 mV from ground potential. If V_{REF} is less than 10V, a finer V_{OS} adjustment is required. It is helpful to increase the resolution of the V_{OS} adjust procedure by connecting a 1 kΩ resistor between the inverting input of the op amp to ground. After V_{OS} has been adjusted, remove the 1 kΩ.

Full-Scale Adjust (*Figure 4*)

Switch high all the digital inputs, A1–A10, and measure the op amp output voltage. Use a 500Ω potentiometer, as shown, to bring $\|V_{OUT}\|$ to a voltage equal to $V_{REF} \times 1023/1024$.

SELECTING AND COMPENSATING THE OPERATIONAL AMPLIFIER

Op Amp Family	C_F	R_i	P	V_W	Circuit Settling Time, t_s	Circuit Small Signal BW
LF357	10 pF	2.4k	25k	V^+	1.5 μs	1M
LF356	22 pF	∞	25k	V^+	3 μs	0.5M
LF351	24 pF	∞	10k	V^-	4 μs	0.5M
LM741	0	∞	10k	V^-	40 μs	200 kHz



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$$V_{OUT} = -V_{REF}\left(\frac{A1}{2} + \frac{A2}{4} + \frac{A3}{8} + \bullet\bullet\bullet\frac{A10}{1024}\right)$$

$$-10V \le V_{REF} \le 10V$$

$$0 \le V_{OUT} \le -\frac{1023}{1024}V_{REF}$$

where $A_N = 1$ if the A_N digital input is high

$A_N = 0$ if the A_N digital input is low

FIGURE 3. Basic Connection: Unipolar or 2-Quadrant Multiplying Configuration (Digital Attenuator)

Typical Applications (Continued)

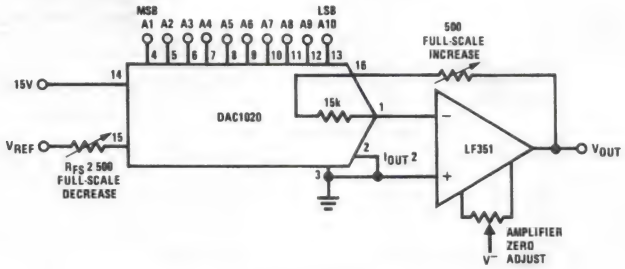


FIGURE 4. Full-Scale Adjust

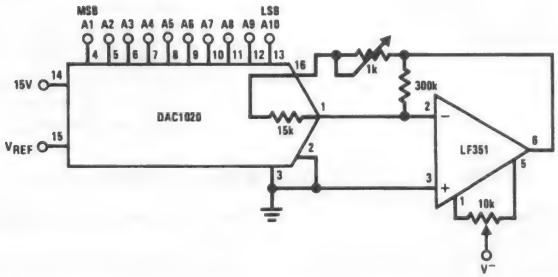
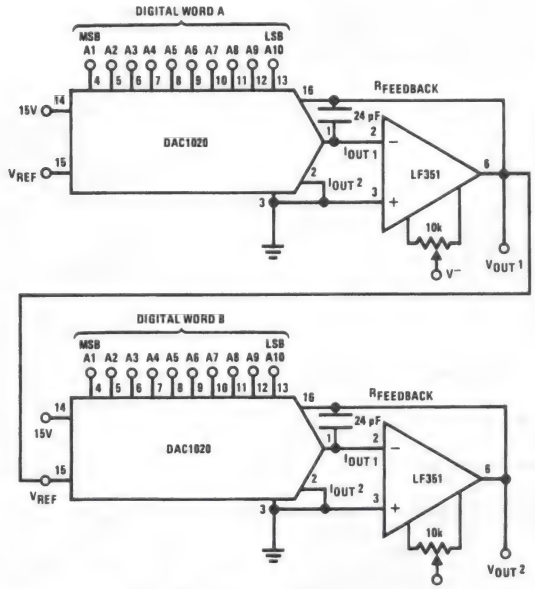


FIGURE 5. Alternate Full-Scale Adjust: (Allows Increasing or Decreasing the Gain)



$$V_{OUT1} = -V_{REF}\left(\frac{A1}{2} + \frac{A2}{4} + \frac{A3}{8} + \cdots \frac{A10}{1024}\right)$$

$$V_{OUT2} = V_{REF}\left(\frac{A1}{2} + \frac{A2}{4} + \frac{A3}{8} + \cdots \frac{A10}{1024}\right) \times \left(\frac{B1}{2} + \frac{B2}{4} + \frac{B3}{8} + \cdots \frac{B10}{1024}\right)$$

where V_{REF} can be an AC signal

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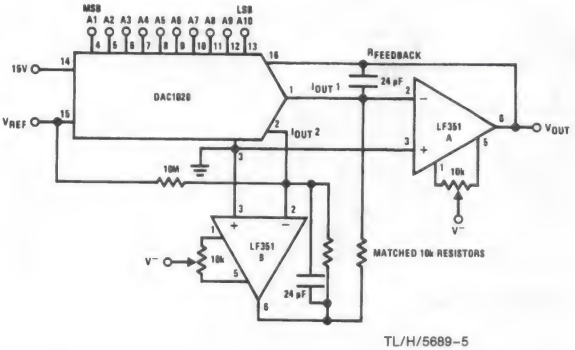
FIGURE 6. Precision Analog-to-Digital Multiplier

Typical Applications (Continued)



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$$V_{OUT} = -V_{REF}\left(\frac{A1}{2} + \frac{A2}{4} + \bullet\bullet\bullet + \frac{A10}{1024} - \frac{1}{1024}\right)$$

where: AN = +1 if A_N input is high
AN = −1 if A_N input is low

COMPLEMENTARY OFFSET BINARY (BIPOLAR) OPERATION

DIGITAL INPUT										V_{OUT}
0	0	0	0	0	0	0	0	0	0	$+V_{REF}$
0	0	0	0	0	0	0	0	0	1	$V_{REF} \times 1022/1024$
0	1	1	1	1	1	1	1	1	1	$V_{REF} \times 2/1024$
1	0	0	0	0	0	0	0	0	0	0
1	0	0	0	0	0	0	0	0	1	$-V_{REF} \times 2/1024$
1	1	1	1	1	1	1	1	1	1	$-V_{REF}\,(1022/1024)$

Note that:

- $I_{OUT\,1} + I_{OUT\,2} = \dfrac{V_{REF}}{R_{LADDER}} \times \left(\dfrac{1023}{1024}\right)$
- By doubling the output range we get half the resolution
- The 10M resistor, adds a 1 LSB "thump", to allow full offset binary operation where the output reaches zero for the half-scale code. If symmetrical output excursions are required, omit the 10M resistor.

FIGURE 7. Bipolar 4-Quadrant Multiplying Configuration

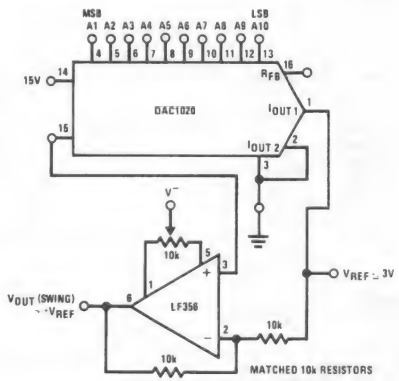
Operational Amplifiers V_{OS} Adjust (*Figure 7*)

a) Switch all the digital inputs high; adjust the V_{OS} potentiometer of op amp B to bring its output to a value equal to $-(V_{REF}/1024)$ (V).

b) Switch the MSB high and the remaining digital inputs low. Adjust the V_{OS} potentiometer of op amp A, to bring its output value to within a 1 mV from ground potential. For $V_{REF} < 10V$, a finer adjust is necessary, as already mentioned in the previous application.

Gain Adjust (Full-Scale Adjust)

Assuming that the external 10k resistors are matched to better than 0.1%, the gain adjust of the circuit is the same with the one previously discussed.

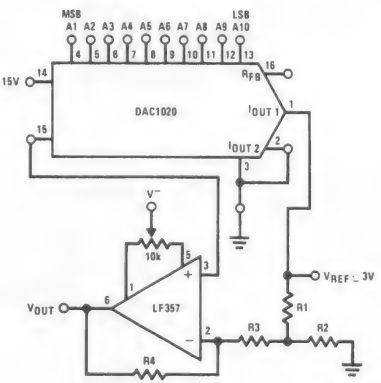


TRUE OFFSET BINARY OPERATION

DIGITAL INPUT										V_{OUT}
1	1	1	1	1	1	1	1	1	1	$V_{REF} \times 1022/1024$
1	0	0	0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0	0	0	$-V_{REF}$

t_s = 1.8 μs
use LM336 for a voltage reference

FIGURE 8. Bipolar Configuration with a Single Op Amp



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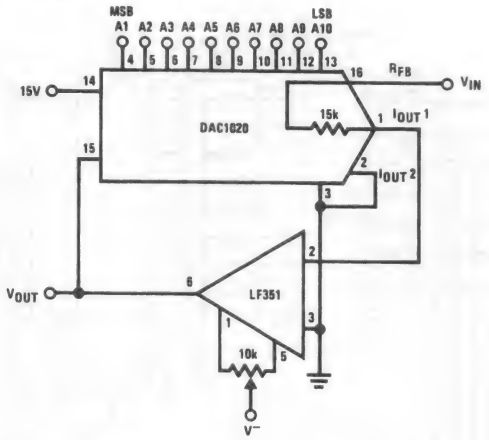
- $R4 = (2A_V^- - 1)\,R, \dfrac{R2}{R1} = \dfrac{A_V^-}{A_V^- - 1}$,

  $R3 + R1 \| R2 = R; A_V^- = \dfrac{V_{OUT(PEAK)}}{V_{REF}}, R = 20k$

- Example: $V_{REF} = 2V$, V_{OUT} (swing) ≅ ±10V: $A_V^- = 5V$
  Then R4 = 9R, R1 = 0.8 R2. If R1 = 0.2R then R2 = 0.25R, R3 = 0.64R

FIGURE 9. Bipolar Configuration with Increased Output Swing

Typical Applications (Continued)

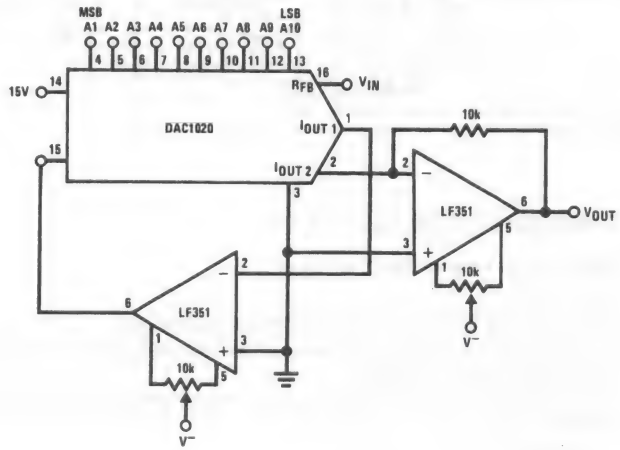


$$V_{OUT} = \frac{-V_{REF}}{\left(\frac{A1}{2} + \frac{A2}{4} + \frac{A3}{8} + \ldots \frac{A10}{1024}\right)}$$

where: V_{REF} can be an AC signal

- By connecting the DAC in the feedback loop of an operational amplifier a linear digitally control gain block can be realized
- Note that with all digital inputs low, the gain of the amplifier is infinity, that is, the op amp will saturate. In other words, we cannot divide the V_{REF} by zero!

FIGURE 10. Analog-to-Digital Divider (or Digitally Gain Controlled Amplifier)



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$$V_{OUT} = V_{REF}\left[\frac{\frac{\overline{A1}}{2} + \frac{\overline{A2}}{4} + \ldots + \frac{\overline{A10}}{1024}}{\frac{A1}{2} + \frac{A2}{4} + \ldots + \frac{A10}{1024}}\right] \text{ or } V_{OUT} = V_{REF}\left(\frac{1023 - N}{N}\right)$$

where: $0 \leq N \leq 1023$

$N = 0$ for A_N = all zeros

$N = 1$ for A10 = 1, A1–A9 = 0

.

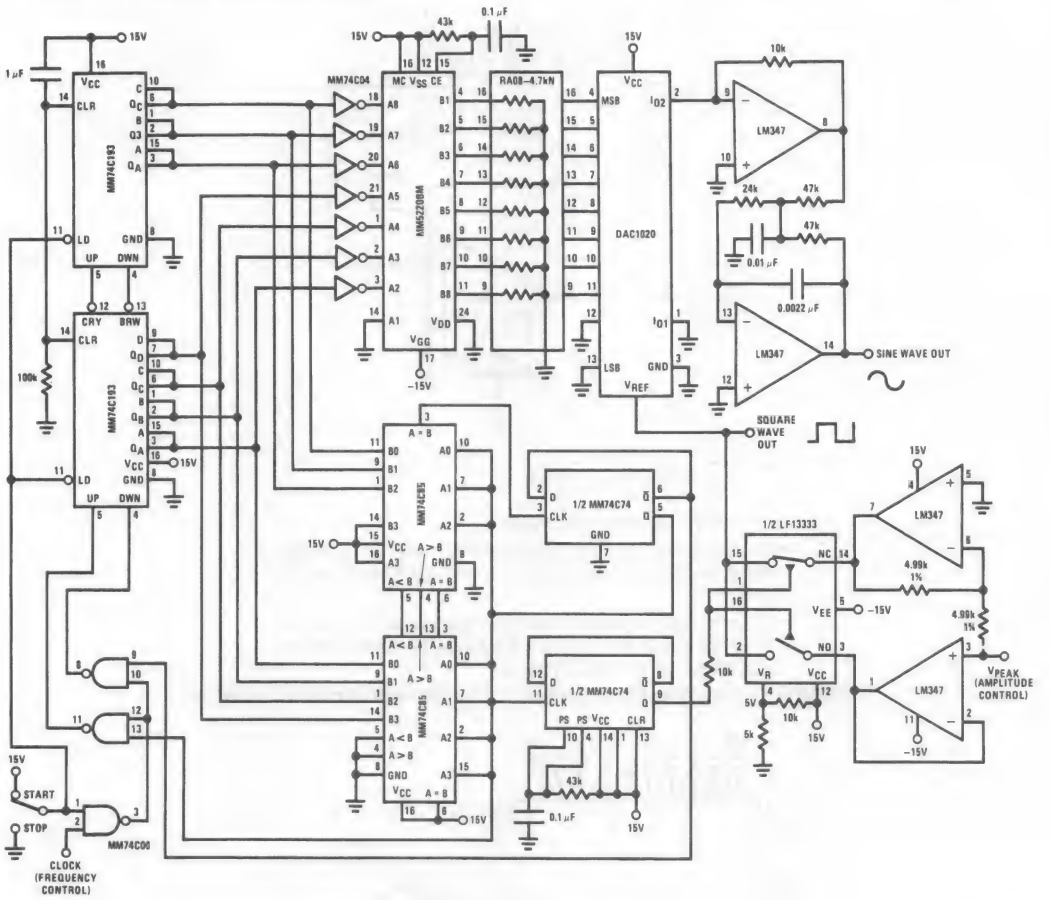
.

.

$N = 1023$ for A_N = all 1's

FIGURE 11. Digitally controlled Amplifier-Attenuator

Typical Applications (Continued)



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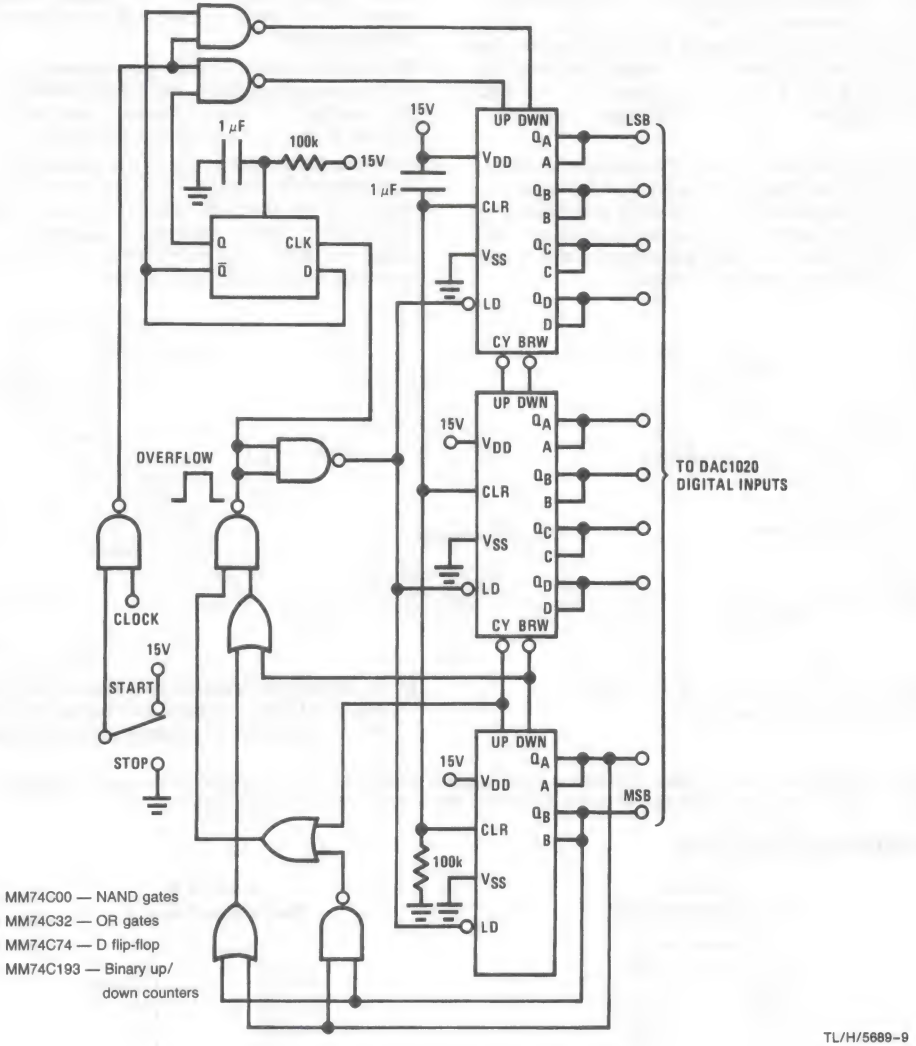
- Output frequency $= \frac{f_{CLK}}{512}$; $f_{MAX} \cong 2$ kHz
- Output voltage range = 0V − 10V peak
- THD < 0.2%
- Excellent amplitude and frequency stability with temperature
- Low pass filter shown has a 1 kHz corner (for output frequencies below 10 Hz, filter corner should be reduced)
- Any periodic function can be implemented by modifying the contents of the look up table ROM
- No start up problems

FIGURE 12. Precision Low Frequency Sine Wave Oscillator Using Sine Look-Up ROM

Typical Applications (Continued)



- Binary up/down counter digitally "ramps" the DAC output
- Can stop counting at any desired 10-bit input code
- Senses up or down count overflow and automatically reverses direction of count

FIGURE 13. A Useful Digital Input Code Generator for DAC Attenuator or Amplifier Circuits

Definition of Terms

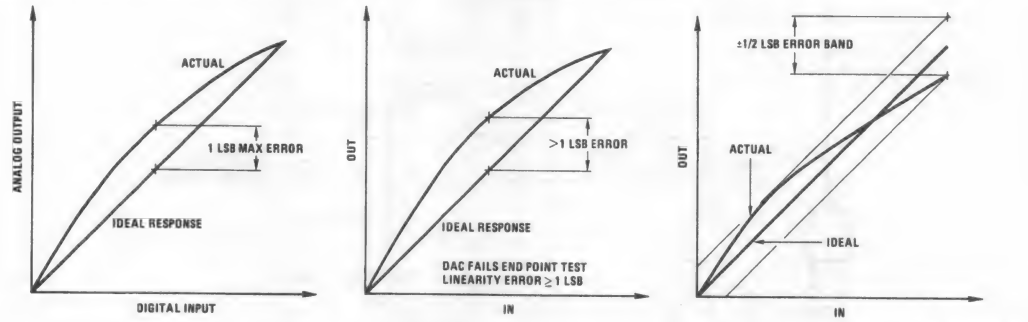
Resolution: Resolution is defined as the reciprocal of the number of discrete steps in the D/A output. It is directly related to the number of switches or bits within the D/A. For example, the DAC1020 has 2^{10} or 1024 steps while the DAC1220 has 2^{12} or 4096 steps. Therefore, the DAC1020 has 10-bit resolution, while the DAC1220 has 12-bit resolution.

Linearity Error: Linearity error is the maximum deviation from *a straight line passing through the endpoints of the D/A transfer characteristic.* It is measured after calibrating for zero (see V_{OS} adjust in typical applications) and full-scale. Linearity error is a design parameter intrinsic to the device and cannot be externally adjusted.

Power Supply Sensitivity: Power supply sensitivity is a measure of the effect of power supply changes on the D/A full-scale output.

Settling Time: Full-scale settling time requires a zero to full-scale or full-scale to zero output change. Settling time is the time required from a code transition until the D/A output reaches within $\pm\frac{1}{2}$ LSB of final output value.

Full-Scale Error: Full-scale error is a measure of the output error between an ideal D/A and the actual device output. Ideally, for the DAC1020 full-scale is $V_{REF}-1$ LSB. For $V_{REF}=10V$ and unipolar operation, $V_{FULL\text{-}SCALE}=10.0000V-9.8\ mV=9.9902V$. Full-scale error is adjustable to zero as shown in *Figure 5*.



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a **b1** **b2**

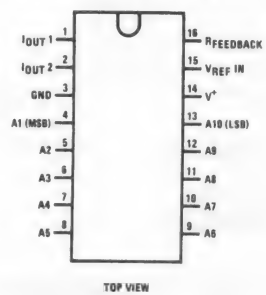
(a) End point test after zero and full-scale adjust. The DAC has 1 LSB linearity error.

(b) By shifting the full-scale calibration on of the DAC of *Figure (b1)* we could pass the "best straight line" (b2) test and meet the $\pm\frac{1}{2}$ linearity error specification.

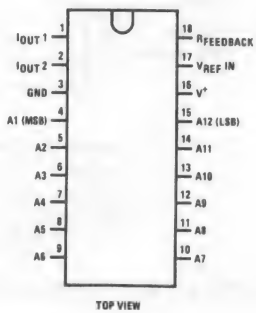
Note. (a), (b1) and (b2) above illustrate the difference between "end point" National's linearity test (a) and "best straight line" test. Note that both devices in (a) and (b2) meet the $\pm\frac{1}{2}$ LSB linearity error specification but the end point test is a more "real life" way of characterizing the DAC.

Connection Diagrams

DAC102X
Dual-In-Line Package



DAC122X
Dual-In-Line Package



TL/H/5689–11

National Semiconductor

MICRO-DAC™ DAC1208/DAC1209/DAC1210/DAC1230/DAC1231/DAC1232 12-Bit, μP Compatible, Double-Buffered D to A Converters

General Description

The DAC1208 and the DAC1230 series are 12-bit multiplying D to A converters designed to interface directly with a wide variety of microprocessors (8080, 8048, 8085, Z-80, etc.). Double buffering input registers and associated control lines allow these DACs to appear as a two-byte "stack" in the system's memory or I/O space with no additional interfacing logic required.

The DAC1208 series provides all 12 input lines to allow single buffering for maximum throughput when used with 16-bit processors. These input lines can also be externally configured to permit an 8-bit data interface. The DAC1230 series can be used with an 8-bit data bus directly as it internally formulates the 12-bit DAC data from its 8 input lines. All of these DACs accept left-justified data from the processor.

The analog section is a precision silicon-chromium (Si-Cr) R-2R ladder network and twelve CMOS current switches. An inverted R-2R ladder structure is used with the binary weighted currents switched between the I_{OUT1} and I_{OUT2} maintaining a constant current in each ladder leg independent of the switch state. Special circuitry provides TTL logic input voltage level compatibility.

The DAC1208 series and DAC1230 series are the 12-bit members of a family of microprocessor compatible DACs (MICRO-DACs™). For applications requiring other resolutions, the DAC1000 series for 10-bit and DAC0830 series for 8-bit are available alternatives.

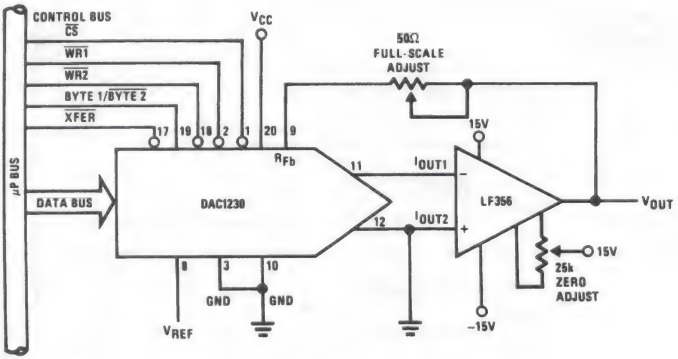
Features

- Linearity specified with zero and full-scale adjust only
- Direct interface to all popular microprocessors
- Double-buffered, single-buffered or flow through digital data inputs
- Logic inputs which meet TTL voltage level specs (1.4V logic threshold)
- Works with ±10V reference—full 4-quadrant multiplication
- Operates stand-alone (without μP) if desired
- All parts guaranteed 12-bit monotonic
- DAC1230 series is pin compatible with the DAC0830 series 8-bit MICRO-DACs

Key Specifications

■ Current Settling Time	1 μs
■ Resolution	12 Bits
■ Linearity (Guaranteed over temperature)	10, 11, or 12 Bits of FS
■ Gain Tempco	1.3 ppm/°C
■ Low Power Dissipation	20 mW
■ Single Power Supply	5 V_{DC} to 15 V_{DC}

Typical Application



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Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.
(Notes 1 and 2)

Supply Voltage (V_{CC})	17 V_{DC}
Voltage at Any Digital Input	V_{CC} to GND
Voltage at V_{REF} Input	±25V
Storage Temperature Range	−65°C to +150°C
Package Dissipation at T_A = 25°C (Note 3)	500 mW
DC Voltage Applied to I_{OUT1} or I_{OUT2} (Note 4)	−100 mV to V_{CC}
ESD Susceptability	800V

Operating Conditions

Lead Temperature (Soldering, 10 seconds)	300°C
Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
DAC1208LCJ, DAC1209LCJ, DAC1210LCJ, DAC1230LCJ, DAC1231LCJ, DAC1232LCJ	$-40°C \le T_A \le +85°C$
DAC1208LCJ-1, DAC1209LCJ-1, DAC1210LCJ-1, DAC1230LCJ-1, DAC1231LCJ-1, DAC1232LCJ-1	$0°C \le T_A \le +70°C$
Range of V_{CC}	4.75 V_{DC} to 16 V_{DC}
Voltage at Any Digital Input	V_{CC} to GND

Electrical Characteristics

V_{REF} = 10.000 V_{DC}, V_{CC} = 11.4 V_{DC} to 15.75 V_{DC} unless otherwise noted. **Boldface limits apply from T_{MIN} to T_{MAX} (see Note 13);** all other limits $T_A = T_J$ = 25°C.

Parameter	Conditions	Notes	Typ (Note 10)	Tested Limit (Note 5)	Design Limit (Note 6)	Units
Resolution			12	**12**	**12**	Bits
Linearity Error (End Point Linearity)	Zero and Full-Scale Adjusted	4, 7, 13				
	DAC1208, DAC1230			**0.012**	**0.012**	% of FSR
	DAC1209, DAC1231			**0.024**	**0.024**	% of FSR
	DAC1210, DAC1232			**0.050**	**0.05**	% of FSR
Differential Non-Linearity	Zero and Full-Scale Adjusted	4, 7, 13				
	DAC1208, DAC1230			**0.018**	**0.018**	% of FSR
	DAC1209, DAC1231			**0.024**	**0.024**	% of FSR
	DAC1210, DAC1232			**0.050**	**0.05**	% of FSR
Monotonicity		4	12	**12**	**12**	Bits
Gain Error (Min)	Using Internal R_{Fb} V_{ref} = ±10V, ±1V	7	−0.1	0.0		% of FSR
Gain Error (Max)		7	−0.1	−0.2		% of FSR
Gain Error Tempco		7	±1.3		**±6.0**	ppm of FS/°C
Power Supply Rejection	All Digital Inputs Latched High	7	±3.0	±30		ppm of FSR/V
Reference Input Resistance (Min)		13	15	**10**	**10**	kΩ
Reference Input Resistance (Max)			15	**20**	**20**	
Output Feedthrough Error	V_{REF} = 20 Vp-p, f = 100 kHz All Data Inputs Latched Low	9	3.0			mVp-p
Output Capacitance	All Data Inputs Latched High I_{OUT1}				200	pF
	All Data Inputs Latched High I_{OUT2}				70	pF
	All Data Inputs Latched Low I_{OUT1}				70	pF
	All Data Inputs Latched Low I_{OUT2}				200	pF
Supply Current Drain		13		2.0	**2.5**	mA
Output Leakage Current I_{OUT1}	All Data Inputs Latched Low	11, 13	0.1	**15**	**15**	nA
I_{OUT2}	All Data Inputs Latched High	11, 13	0.1	**15**	**15**	nA
Digital Input Threshold	Low Threshold	13		**0.8**	**0.8**	V_{DC}
	High Threshold	13		**2.2**	**2.2**	V_{DC}
Digital Input Currents	Digital Inputs <0.8V	13		**−200**	**−200**	μA_{DC}
	Digital Inputs >2.2V	13		**10**	**10**	μA_{DC}

Electrical Characteristics (Continued)

V_{REF} = 10.000 V_{DC}, V_{CC} = 11.4 V_{DC} to 15.75 V_{DC} unless otherwise noted. **Boldface limits apply from T_{MIN} to T_{MAX} (see Note 13);** all other limits $T_A = T_J$ = 25°C.

Symbol	Parameter	Conditions	See Note	Typ (Note 10)	Tested Limit (Note 5)	Design Limit (Note 6)	Units
AC CHARACTERISTICS							
t_s	Current Setting Time	V_{IL} = 0V, V_{IH} = 5V		1.0			μs
t_W	Write and XFER Pulse Width Min.	V_{IL} = 0V, V_{IH} = 5V	8	50		320 **320**	ns
t_{DS}	Data Setup Time Min.	V_{IL} = 0V, V_{IH} = 5V		70		320 **320**	
t_{DH}	Data Hold Time Min.	V_{IL} = 0V, V_{IH} = 5V		30		90 **90**	
t_{CS}	Control Setup Time Min.	V_{IL} = 0V, V_{IH} = 5V		60		320 **320**	
t_{CH}	Control Hold Time Min.	V_{IL} = 0V, V_{IH} = 5V		0		10	

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to GND, unless otherwise specified.

Note 3: This 500 mW specification applies for all packages. The low intrinsic power dissipation of this part (and the fact that there is no way to significantly modify the power dissipation) removes concern for heat sinking.

Note 4: Both I_{OUT1} and I_{OUT2} must go to ground or the virtual ground of an operational amplifier. The linearity error is degraded by approximately $V_{OS} \div V_{REF}$. For example, if V_{REF} = 10V then a 1 mV offset, V_{OS}, on I_{OUT1} or I_{OUT2} will introduce an additional 0.01% linearity error.

Note 5: Tested and guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 6: Design limits are guaranteed but not 100% tested. These limits are not used to calculate outgoing quality levels. Guaranteed for V_{CC} = 11.4V to 15.75V and V_{REF} = −10V to +10V.

Note 7: The unit FSR stands for full-scale range. Linearity Error and Power Supply Rejection specs are based on this unit to eliminate dependence on a particular V_{REF} value to indicate the true performance of the part. The Linearity Error specification of the DAC1208 is 0.012% of FSR(max). This guarantees that after performing a zero and full-scale adjustment, the plot of the 4096 analog voltage outputs will each be within 0.012% × V_{REF} of a straight line which passes through zero and full-scale. The unit ppm of FSR(parts per million of full-scale range) and ppm of FS(parts per million of full-scale) are used for convenience to define specs of very small percentage values, typical of higher accuracy converters. In this instance, 1 ppm of FSR = $V_{REF}/10^6$ is the conversion factor to provide an actual output voltage quantity. For example, the gain error tempco spec of ±6 ppm of FS/°C represents a worst-case full-scale gain error change with temperature from −40°C to +85°C of ±(6)($V_{REF}/10^6$)(125°C) or ±0.75 (10^{-3}) V_{REF} which is ±0.075% of V_{REF}.

Note 8: This spec implies that all parts are guaranteed to operate with a write pulse or transfer pulse width (t_W) of 320 ns. A typical part will operate with t_W of only 100 ns. The entire write pulse must occur within the valid data interval for the specified t_W, t_{DS}, t_{DH} and t_S to apply.

Note 9: To achieve this low feedthrough in the D package, the user must ground the metal lid. If the lid is left floating the feedthrough is typically 6 mV.

Note 10: Typicals are at 25°C and represent the most likely parametric norm.

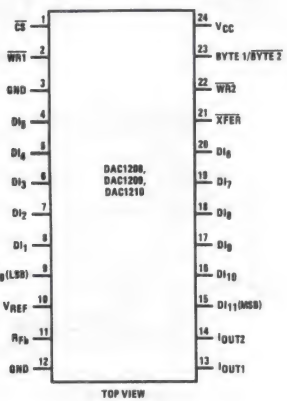
Note 11: A 10 nA leakage current with R_{Fb} = 20k and V_{REF} = 10V corresponds to a zero error of $(10 \times 10^{-9} \times 20 \times 10^3) \times 100\%$ 10V or 0.002% of FS.

Note 12: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Note 13: Tested limit for −1 suffix parts applies only at 25°C.

Connection Diagrams

Dual-In-Line Package



TOP VIEW

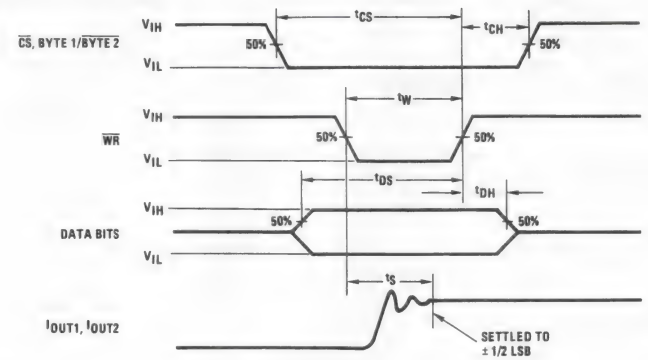
Dual-In-Line Package

TOP VIEW

TL/H/5690–2

See Ordering Information

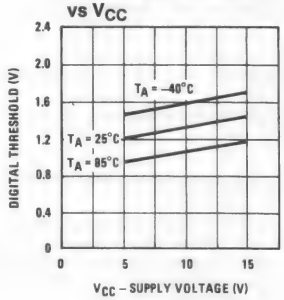
Switching Waveforms



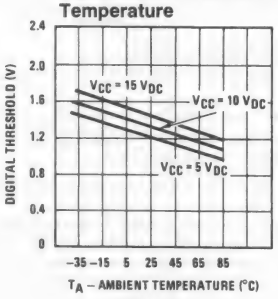
TL/H/5690–3

Typical Performance Characteristics

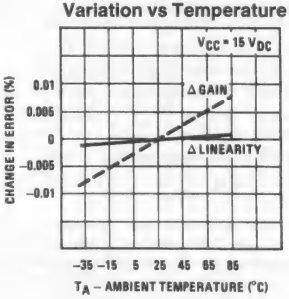
Digital Input Threshold vs V_{CC}



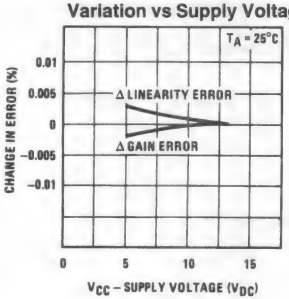
Digital Input Threshold vs Temperature



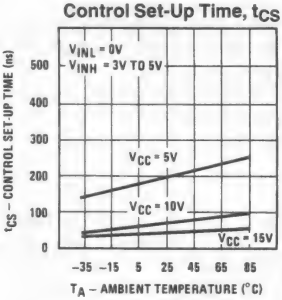
Gain and Linearity Error Variation vs Temperature



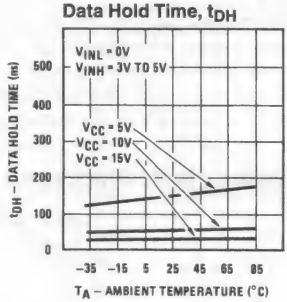
Gain and Linearity Error Variation vs Supply Voltage



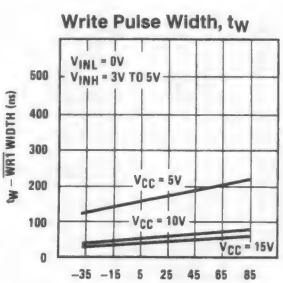
Control Set-Up Time, t_{CS}



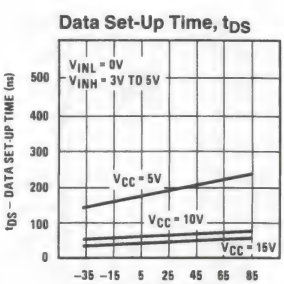
Data Hold Time, t_{DH}



Write Pulse Width, t_W



Data Set-Up Time, t_{DS}



TL/H/5690–4

Definition of Package Pinouts

CONTROL SIGNALS (all control signals are level actuated)

$\overline{CS}$: Chip Select (active low). The $\overline{CS}$ will enable $\overline{WR1}$.

$\overline{WR1}$: Write 1. The active low $\overline{WR1}$ is used to load the digital data bits (DI) into the input latch. The data in the input latch is latched when $\overline{WR1}$ is high. The 12-bit input latch is split into two latches. One holds the first 8 bits, while the other holds 4 bits. The Byte 1/$\overline{Byte\ 2}$ control pin is used to select both latches when Byte 1/$\overline{Byte\ 2}$ is high or to overwrite the 4-bit input latch when in the low state.

Byte 1/$\overline{Byte\ 2}$: Byte Sequence Control. When this control is high, all 12 locations of the input latch are enabled. When low, only the four least significant locations of the input latch are enabled.

$\overline{WR2}$: Write 2 (active low). The $\overline{WR2}$ will enable $\overline{XFER}$.

$\overline{XFER}$: Transfer Control Signal (active low). This signal, in combination with $\overline{WR2}$, causes the 12-bit data which is available in the input latches to transfer to the DAC register.

DI_0 to DI_{11}: Digital Inputs. DI_0 is the least significant digital input (LSB) and DI_{11} is the most significant digital input (MSB).

I_{OUT1}: DAC Current Output 1. I_{OUT1} is a maximum for a digital code of all 1s in the DAC register, and is zero for all 0s in the DAC register.

I_{OUT2}: DAC Current Output 2. I_{OUT2} is a constant minus I_{OUT1}, or $I_{OUT1} + I_{OUT2} = \text{constant}$ (for a fixed reference voltage). This constant current is

$$V_{REF} \times \left(1 - \frac{1}{4096}\right)$$

divided by the reference input resistance.

R_{FB}: Feedback Resistor. The feedback resistor is provided on the IC chip for use as the shunt feedback resistor for the external op amp which is used to provide an output voltage for the DAC. This on-chip resistor should always be used (not an external resistor) since it matches the resistors in the on-chip R-2R ladder and tracks these resistors over temperature.

V_{REF}: Reference Voltage Input. This input connects an external precision voltage source to the internal R-2R ladder. V_{REF} can be selected over the range of 10V to −10V. This is also the analog voltage input for a 4-quadrant multiplying DAC application.

V_{CC}: Digital Supply Voltage. This is the power supply pin for the part. V_{CC} can be from 5 V_{DC} to 15 V_{DC}. Operation is optimum for 15 V_{DC}.

GND: Pins 3 and 12 of the DAC1208, DAC1209, and DAC1210 must be connected to ground. Pins 3 and 10 of the DAC1230, DAC1231, and DAC1232 must be connected to ground. It is important that I_{OUT_1} and I_{OUT_2} are at ground potential for current switching applications. Any difference of potential (V_{OS} on these pins) will result in a linearity change of

$$\frac{V_{OS}}{3\ V_{REF}}$$

For example, if V_{REF} = 10V and these ground pins are 9 mV offset from I_{OUT_1} and I_{OUT_2}, the linearity change will be 0.03%.

Definition of Terms

Resolution: Resolution is defined as the reciprocal of the number of discrete steps in the DAC output. It is directly related to the number of switches or bits within the DAC. For example, the DAC1208 has 2^{12} or 4096 steps and therefore has 12-bit resolution.

Linearity Error: Linearity error is the maximum deviation from a *straight line passing through the endpoints of the DAC transfer characteristic.* It is measured after adjusting for zero and full-scale. Linearity error is a parameter intrinsic to the device and cannot be externally adjusted.

National's linearity test (a) and the best straight line test (b) used by other suppliers are illustrated below. The best straight line (b) requires a special zero and FS adjustment for each part, which is almost impossible for the user to determine. The end point test uses a standard zero FS adjustment procedure and is a much more stringent test for DAC linearity.

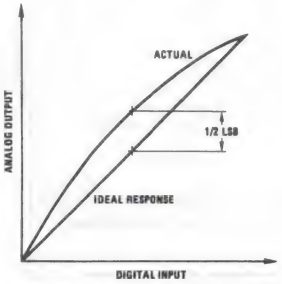
Power Supply Sensitivity: Power supply sensitivity is a measure of the effect of power supply changes on the DAC full-scale output.

Settling Time: Full-scale current settling time requires zero to full-scale or full-scale to zero output change. Settling time is the time required from a code transition until the DAC output reaches within ±½ LSB of the final output value.

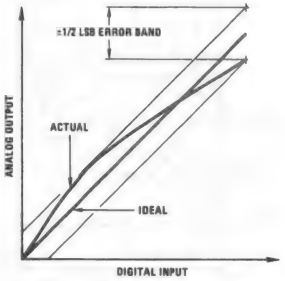
Full-Scale Error: Full-scale error is a measure of the output error between an ideal DAC and the actual device output. Ideally, for the DAC1208 or DAC1230 series, full-scale is $V_{REF} - 1$ LSB. For $V_{REF} = 10V$ and unipolar operation, $V_{FULL\text{-}SCALE} = 10.0000V - 2.44\ mV = 9.9976V$. Full-scale error is adjustable to zero.

Differential Non-Linearity: The difference between any two consecutive codes in the transfer curve from the theoretical 1 LSB is differential non-linearity.

Monotonic: If the output of a DAC increases for increasing digital input code, then the DAC is monotonic. A 12-bit DAC which is monotonic to 12 bits simply means that input increasing digital input codes will produce an increasing analog output.



a) End Point Test After Zero and FS Adjust



b) Shifting FS Adjust to Pass Best Straight Line Test

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4

Application Hints

1.0 DIGITAL INTERFACE

These DACs are designed to provide all of the necessary digital input circuitry to permit a direct interface to a wide variety of microprocessor systems. The timing and logic level convention of the input control signals allow the DACs to be treated as a typical memory device or I/O peripheral with no external logic required in most systems. Essentially these DACs can be mapped as a two-byte stack in memory (or I/O space) to receive their 12 bits of input data in two successive 8-bit data writing sequences. The DAC1230 series is intended for use in systems with an 8-bit data bus. The DAC1208 series provides all 12 digital input lines which can be externally configured to be controlled from an 8-bit bus or can be driven directly from a 16-bit data bus.

All of the digital inputs to these DACs contain a unique threshold regulator circuit to maintain TTL voltage level compatibility independent of the applied V_{CC} to the DAC. Any input can also be driven from higher voltage CMOS logic levels in non-microprocessor based systems. To prevent damage to the chip from static discharge, all unused digital inputs should be tied to V_{CC} or ground. As a trouble-shooting aid, if any digital input is inadvertently left floating, the DAC will interpret the pin as a logic "1".

Double buffered digital inputs allow the DAC to internally format the 12-bit word used to set the current switching R-2R ladder network (see section 2.0) from two 8-bit data write cycles. *Figures 1* and *2* show the internal data registers and their controlling logic circuitry. The timing diagrams for updating the DAC output are shown in sections 1.1, 1.2 and 1.3 for three possible control modes. The method used depends strictly upon the particular application.

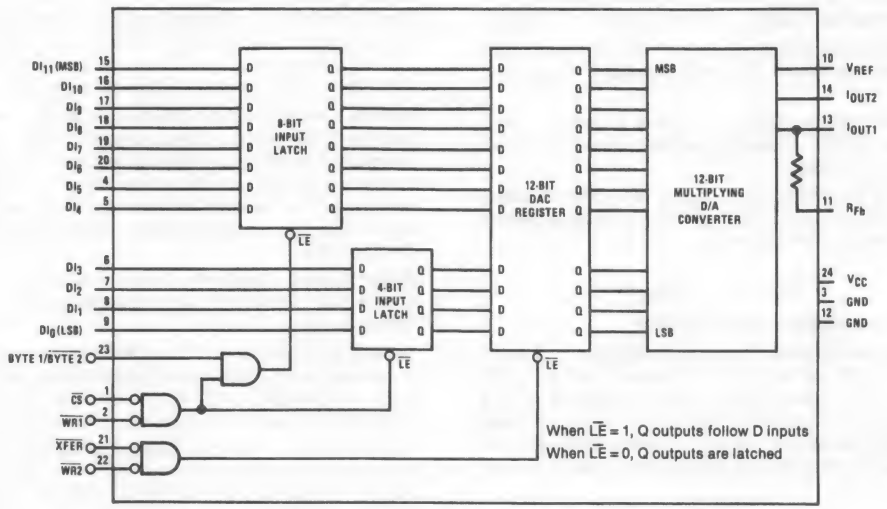


FIGURE 1. DAC1208, DAC1209, DAC1210 Functional Diagram

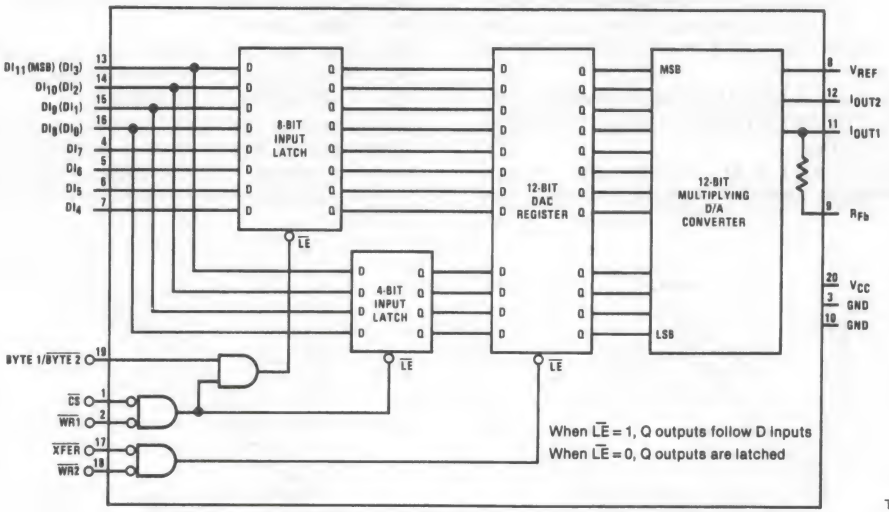
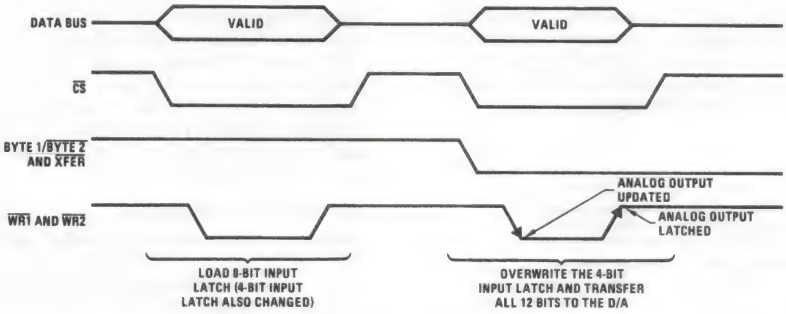


FIGURE 2. DAC1230, DAC1231, DAC1232 Functional Diagram

Application Hints (Continued)

1.1 Automatic Transfer

The 12-bit DAC word is automatically transferred to the DAC register and the R-2R ladder when the second write (the 4 LSBs of the data) occurs.


DATA BUS
VALID
VALID
CS
BYTE 1/BYTE 2 AND XFER
WR1 AND WR2
ANALOG OUTPUT UPDATED
ANALOG OUTPUT LATCHED
LOAD 8-BIT INPUT LATCH (4-BIT INPUT LATCH ALSO CHANGED)
OVERWRITE THE 4-BIT INPUT LATCH AND TRANSFER ALL 12 BITS TO THE D/A

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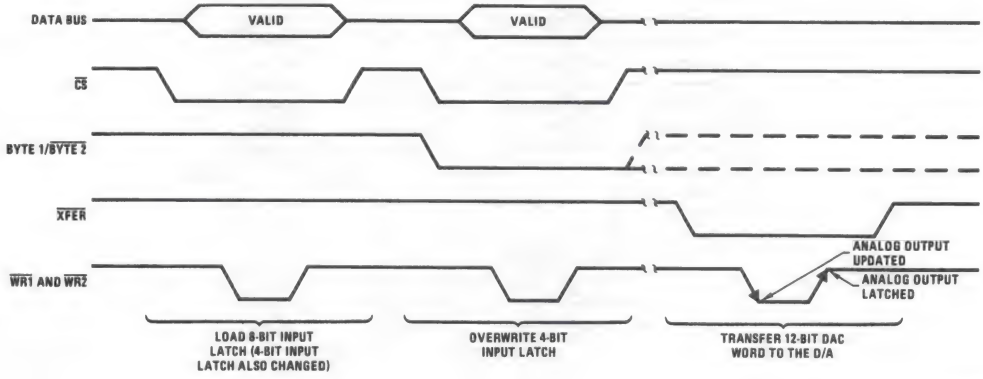
1.2 Independent Processor Transfer Control

In this case a separate address is decoded to provide the $\overline{XFER}$ signal. This allows the processor to load the next required DAC word but not change the analog output until some time later, most useful for the simultaneous updating of several DACs in a system where their $\overline{XFER}$ lines would be tied together.



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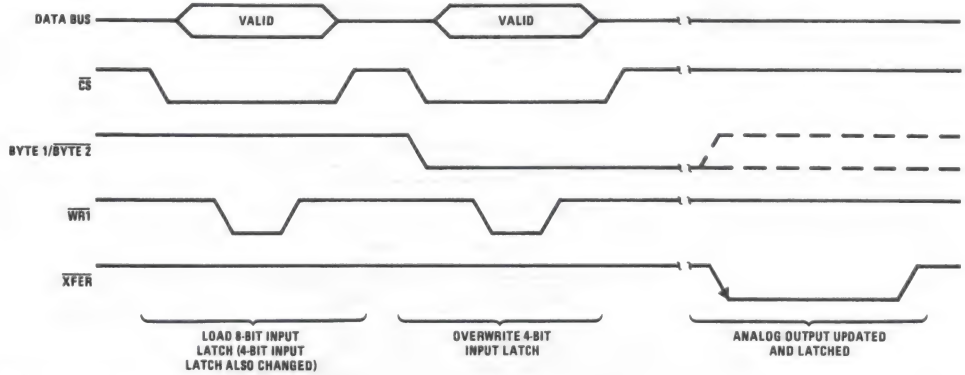
1.3 Transfer via an External Strobe

This method is basically the same as the previous operation except the $\overline{XFER}$ signal is provided by a device other than the processor. This allows the DAC to hold the code for a conditional analog output signal which will be required on demand from an external monitoring device (an analog voltage comparator for instance).



$\overline{WR2}$ tied to a logic low (0V)

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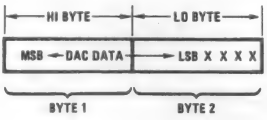
Application Hints (Continued)

1.4 Left-Justified Data Format

It is important to realize that the input registers of these DACs are arranged to accept a left-justified data word from the microprocessor with the most significant 8 bits coming first (Byte 1) and the lower 4 bits second. Left justification simply means that the binary point is assumed to be located to the left of the most significant bit. *Figure 3* shows how the 12 bits of DAC data should be arranged in 2 8-bit registers of an 8-bit processor before being written to the DAC.



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X = don't care

FIGURE 3. Left-Justified Data Format

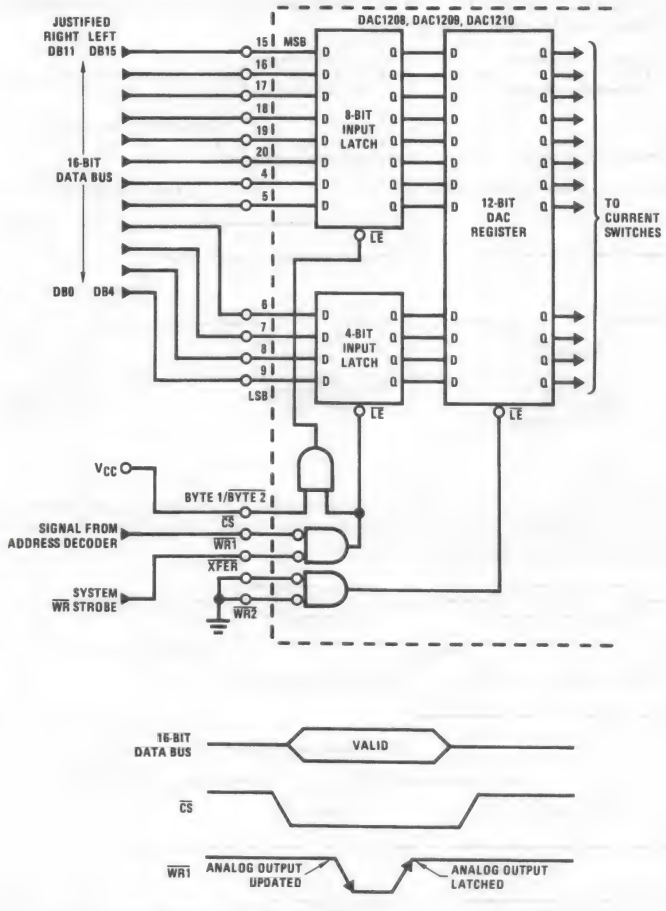
1.5 16-Bit Data Bus Interface

The DAC1208 series provides all 12 digital input lines to permit a direct parallel interface to a 16-bit data bus. In this instance, double buffering is not always necessary (unless a simultaneous updating of several DACs or a data transfer via an external strobe is desired) so the 12-bit DAC register can be wired to flow-through whereby its Q outputs always reflect the state of its D inputs. The external connections required and the timing diagram for this single buffered application are shown in *Figure 4*. Note that either left or right-justified data from the processor can be accommodated with a 16-bit data bus.

1.6 Flow-Through Operation

Through primarily designed to provide microprocessor interface compatibility, the MICRO-DACs can easily be configured to allow the analog output to continuously reflect the state of an applied digital input. This is most useful in appli-

Interface Timing



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XFER and WR2 grounded; Byte 1/Byte 2 tied to V_{CC}.

FIGURE 4. 16-Bit Data Bus Interface for the DAC1208 Series

Application Hints (Continued)

cations where the DAC is used in a continuous feedback control loop and is driven by a binary up/down counter, or in function generation circuits where a ROM is continuously providing DAC data.

Only the DAC1208, DAC1209, DAC1210 devices can have all 12 inputs flow-through. Simply grounding $\overline{CS}$, $\overline{WR1}$, $\overline{WR2}$ and $\overline{XFER}$ and tying Byte 1/$\overline{Byte\ 2}$ high allows both internal registers to follow the applied digital inputs (flow-through) and directly affect the DAC analog output.

1.7 Address Decoding Tips

It is possible to map the MICRO-DACs into system ROM space to allow more efficient use of existing address decoding hardware. The DAC in effect can share the same addresses of any number of ROM locations. The ROM outputs will only be enabled by a READ of its address (gated by the system READ strobe) and the DAC will only accept data that is written to the same address (gated by the system WRITE strobe).

The Byte 1/$\overline{Byte\ 2}$ control function can easily be generated by the processor's least significant address bit (A0) by placing the DAC at two consecutive address locations and utilizing double-byte WRITE instructions which automatically increment or decrement the address. The $\overline{CS}$ and $\overline{XFER}$ signals can then be decoded from the remaining address bits. Care must be taken in selecting the actual address used for Byte 1 of the DAC to prevent a carry (as a result of incrementing the address for Byte 2) from propagating through the address word and changing any of the bits decoded for $\overline{CS}$ or $\overline{XFER}$. *Figure 5* shows how to prevent this effect.

The same problem can occur from a borrow when an autodecremented address is used; but only if the processor's address outputs are inverted before being decoded.

1.8 Control Signal Timing

When interfacing these MICRO-DACs to any microprocessor, there are two important time relationships that must be considered to insure proper operation. The first is the minimum $\overline{WR}$ strobe pulse width which is specified as 320 ns for V_{CC} = 11.4V to 15.75V and operation over temperature, but typically a pulse width of only 250 ns is adequate. A second consideration is that the guaranteed minimum data hold time of 90 ns should be met or erroneous data can be latched. This hold time is defined as the length of time data must be held valid on the digital inputs *after* a qualified (via $\overline{CS}$) $\overline{WR}$ strobe makes a low to high transition to latch the applied data.

If the controlling device or system does not inherently meet these timing specs the DAC can be treated as a slow memory or peripheral and utilize a technique to extend the write strobe. A simple extension of the write time, by adding a wait state, can simultaneously hold the write strobe active and data valid on the bus to satisfy the minimum $\overline{WR}$ pulse

Write Cycle	Address Bits			
	15	2	1*	0**
First (Byte 1)	Decoded to Address DAC		0	1
Second (Byte 2)			1	0

*Starting with a 0 prevents a carry on address incrementing.

**Used as Byte 1/$\overline{Byte2}$ Control.

FIGURE 5

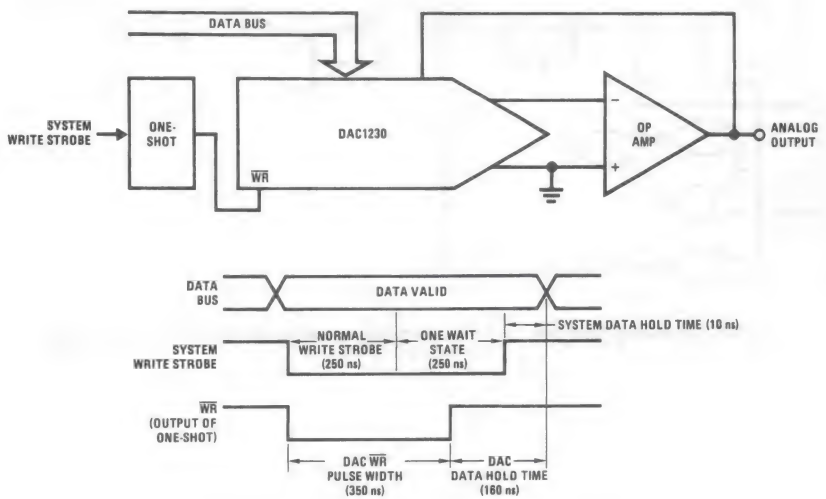

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FIGURE 6. Accommodating a High Speed System

Application Hints (Continued)

width. If this does not provide a sufficient data hold time at the end of the write cycle, a negative edge triggered one-shot can be included between the system write strobe and the $\overline{WR}$ pin of the DAC. This is illustrated in *Figure 6* for an exemplary system which provides a 250 ns $\overline{WR}$ strobe time with a data hold time of only 10 ns.

The proper data set-up time prior to the latching edge (low to high transition) of the $\overline{WR}$ strobe, is insured if the $\overline{WR}$ pulse width is within spec and the data is valid on the bus for the duration of the DAC $\overline{WR}$ strobe.

1.9 Digital Signal Feedthrough

A typical microprocessor is a tremendous potential source of high frequency noise which can be coupled to sensitive analog circuitry. The fast edges of the data and address bus signals generate frequency components of 10's of megahertz and may cause fast transients to appear at the DAC output, even when data is latched internally.

In low frequency or DC applications, low pass filtering can reduce the magnitude of any fast transients. This is most easily accomplished by over-compensating the DAC output amplifier by increasing the value of its feedback capacitor.

In applications requiring a fast output response from the DAC and op amp, filtering may not be feasible. In this event, digital signals can be completely isolated from the DAC circuitry, by the use of a DM74LS374 latch, until a valid $\overline{CS}$ signal is applied to update the DAC. This is shown in *Figure 7*.

A single TRI-STATE® data buffer such as the DM81LS95 can be used to isolate any number of DACs in a system. *Figure 8* shows this isolating circuitry and decoding hardware for a multiple DAC analog output card. Pull-up resistors are used on the buffer outputs to limit the impedance at the DAC digital inputs when the card is not selected. A unique feature of this card is that the DAC $\overline{XFER}$ strobes are controlled by the data bus. This allows a very flexible update of any combination of analog outputs via a transfer word which would contain a zero in the bit position assigned to any of the DACs required to change to a new output value.

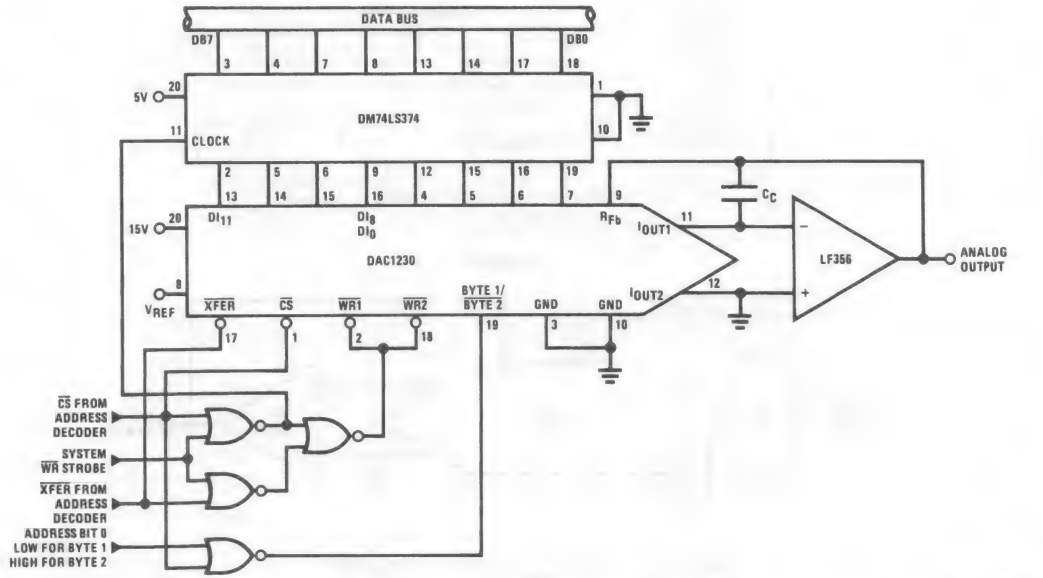

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FIGURE 7. Isolating Data Bus from DAC Circuitry to Eliminate Digital Noise Coupling

Application Hints (Continued)

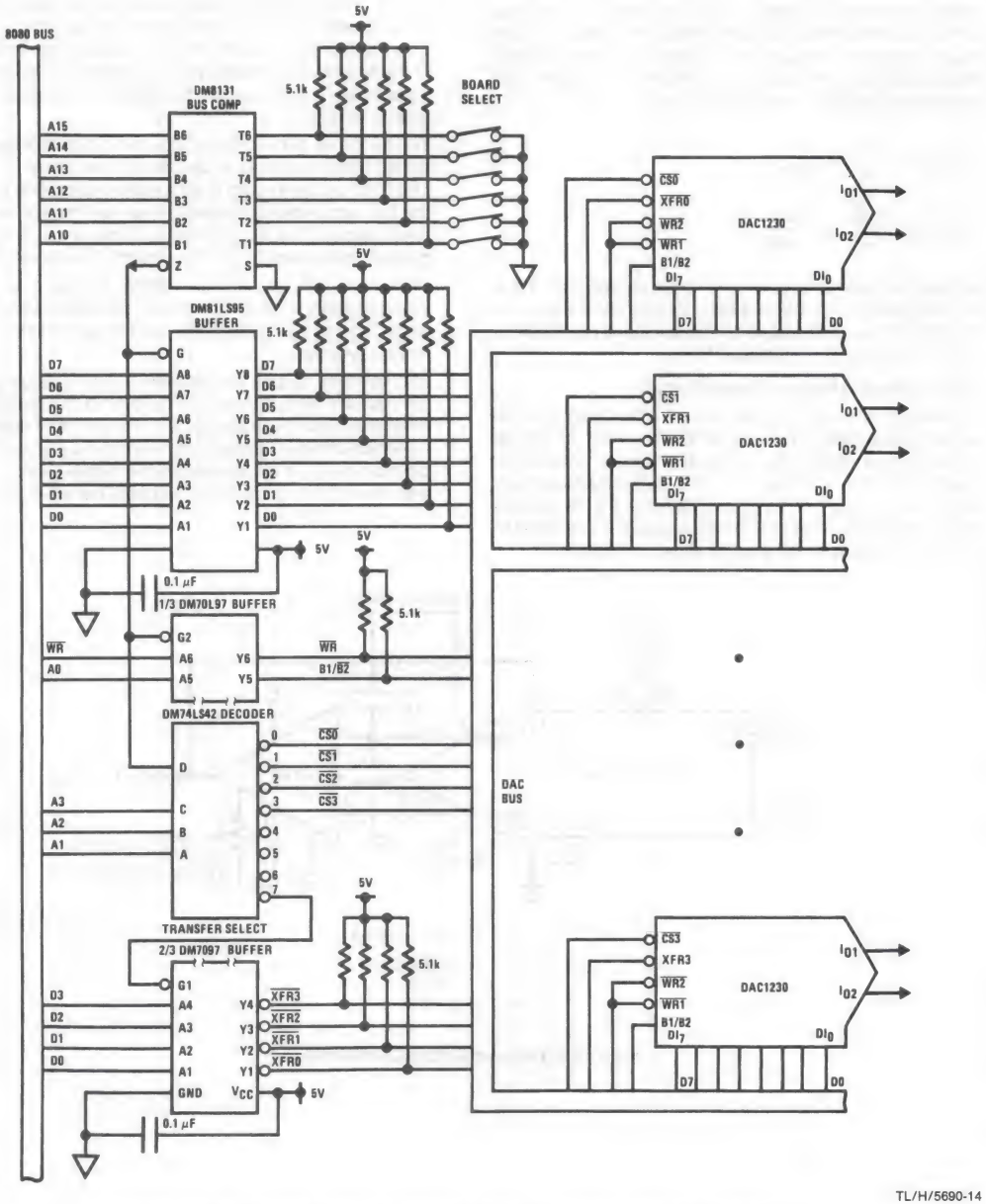

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FIGURE 8. TRI-STATE® Buffers Isolate the Data and Control Lines from the DACs. A Transfer Word Provides a Flexible Update.

Application Hints (Continued)

2.0 ANALOG APPLICATIONS

The analog output signal for these DACs is derived from a conventional R-2R current switching ladder network. A detailed description of this network can be found on the DAC1000 series data sheet. Basically, output I_{OUT1} provides a current directly proportional to the product of the applied reference voltage and the digital input word. A second output, I_{OUT2} will be a current proportional to the complement of the digital input. Specifically:

$$I_{OUT1} = \frac{V_{REF}}{R} \times \frac{D}{4096};$$

$$I_{OUT2} = \frac{V_{REF}}{R} \times \frac{4095 - D}{4096}$$

where D is the decimal equivalent of the applied 12-bit binary word (ranging from 0 to 4095), V_{REF} is the voltage applied to the V_{REF} terminal and R is the internal resistance of the R-2R ladder. R is nominally 15 kΩ.

2.1 Obtaining a Unipolar Output Voltage

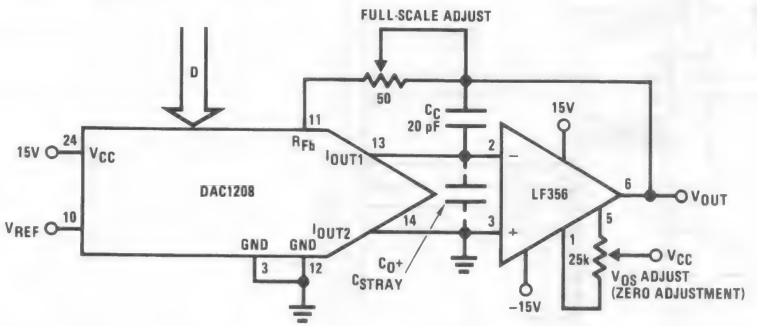
To maintain linearity of output current with changes in the applied digital code, it is important that the voltages at both of the current output pins be as near ground potential (0 V_{DC}) as possible. With $V_{REF} = +10V$ every millivolt appearing at either I_{OUT1} or I_{OUT2} will cause a 0.01% linearity error. In most applications this output current is converted to a voltage by using an op amp as shown in *Figure 9*.

The inverting input of the op amp is a virtual ground created by the feedback from its output through the internal 15 kΩ resistor, R_{Fb}. All of the output current (determined by the digital input and the reference voltage) will flow through R_{Fb} to the output of the amplifier. Two-quadrant operation can be obtained by reversing the polarity of V_{REF} thus causing I_{OUT1} to flow into the DAC and be sourced from the output of the amplifier. The output voltage, in either case, is always equal to $I_{OUT1} \times R_{Fb}$ and is the opposite polarity of the reference voltage.

The reference can be either a stable DC voltage source or an AC signal anywhere in the range from −10V to +10V. The DAC can be thought of as a digitally controlled attenuator: the output voltage is always less than the applied reference voltage. The V_{REF} terminal of the device presents a nominal impedance of 15 kΩ to ground to external circuitry.

Always use the internal R_{Fb} resistor to create an output voltage since this resistor matches (and tracks with temperature) the value of the resistors used to generate the output current (I_{OUT1}).

The selected op amp should have as low a value of input bias current as possible. The product of the bias current times the feedback resistance creates an output voltage error which can be significant in low reference voltage applications. BI-FET™ op amps are highly recommended for use with these DACs because of their very low input current.



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$$V_{OUT} = -(I_{OUT1} \times R_{Fb})$$

$$= \frac{-V_{REF}(D)}{4096}$$

for $0 \leq D \leq 4095$

FIGURE 9. Unipolar Output Configuration

Application Hints (Continued)

Transient response and settling time of the op amp are important in fast data throughput applications. The largest stability problem is the feedback pole created by the feedback resistance, R_{Fb}, and the output capacitance of the DAC. This appears from the op amp output to the (−) input and includes the stray capacitance at this node. Addition of a lead capacitance, C_C in *Figure 9*, greatly reduces overshoot and ringing at the output for a step change in DAC output current.

2.1.1 Zero and Full-Scale Adjustments

For accurate conversions, the input offset voltage of the output amplifier must always be nulled. Amplifier offset errors create an overall degradation of DAC linearity.

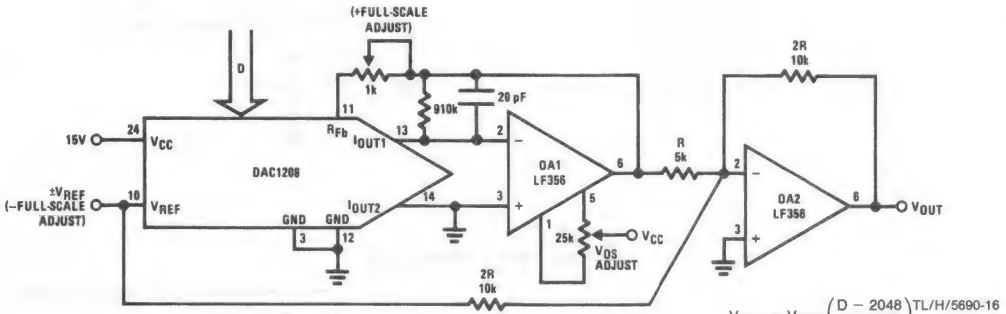
The fundamental purpose of zeroing is to make the voltage appearing at the DAC outputs as near 0 V_{DC} as possible. This is accomplished by shorting out R_{Fb}, the amplifier feedback resistor, and adjusting the v_{OS} nulling potentiometer of the op amp until the output reads zero volts. This is done, of course, with an applied digital code of all zeros if I_{OUT1} is driving the op amp (all ones for I_{OUT2}). The short around R_{Fb} is then removed and the converter is zero adjusted.

A unique feature of this series of DACs is that the full-scale or gain error is guaranteed to be negative. The gain error specification is a measure of how close the value of the internal feedback resistor, R_{Fb}, matches the R-2R ladder resistors. A negative gain error indicates that R_{Fb} is a smaller resistance value than it should be. To adjust this gain error, some resistance must always be added in series with R_{Fb}. The 50Ω potentiometer shown is sufficient to adjust the worst-case gain error for these devices.

2.2 Bipolar Output Voltage from a Fixed Reference

The addition of a second op amp to the unipolar circuit can generate a bipolar output voltage from a fixed reference voltage. This, in effect, gives sign significance to the MSB of the digital input word to allow two quadrant multiplication of the reference voltage. The polarity of the reference can also be reversed to realize full 4-quadrant multiplication. This circuit is shown in *Figure 10*.

This configuration features several improvements over existing circuits for a bipolar output shown with other multiplying DACs. Only the offset voltage of amplifier 1 affects the linearity of the DAC. The offset voltage error of the second op amp (although a constant output error) has no effect on linearity. In addition, this configuration offers a non-interactive positive and negative full-scale calibration procedure.



$$V_{OUT} = V_{REF}\left(\frac{D - 2048}{2048}\right)$$ TL/H/5690-16

for $0 \le D \le 4095$

$$1\ LSB = \frac{|V_{REF}|}{2048}$$

Input Code MSB.....LSB	Ideal V_{OUT}	
	$+V_{REF}$	$-V_{REF}$
1 1 1 1 1 1 1 1 1 1 1 1	$V_{REF} - 1$ LSB	$-\lvert V_{REF}\rvert + 1$ LSB
1 1 0 0 0 0 0 0 0 0 0 0	$V_{REF}/2$	$-\lvert V_{REF}\rvert/2$
1 0 0 0 0 0 0 0 0 0 0 0	0	0
0 1 1 1 1 1 1 1 1 1 1 1	−1 LSB	+1 LSB
0 0 1 1 1 1 1 1 1 1 1 1	$-\frac{V_{REF}}{2} - 1$ LSB	$\frac{\lvert V_{REF}\rvert}{2} + 1$ LSB
0 0 0 0 0 0 0 0 0 0 0 0	$-V_{REF}$	$+\lvert V_{REF}\rvert$

FIGURE 10. Bipolar Output Voltage Configuration

Application Hints (Continued)

2.2.1 Zero and Full-Scale Adjustments

To calibrate the bipolar output circuit, three adjustments are required. The first step is to set all of the digital inputs LOW (to force I_{OUT1} to 0) then null the V_{OS} of amplifier 1 by setting the voltage at its inverting input (pin 2) to zero volts. Next, with a code of all zeros still applied, adjust "−full-scale adjust", the reference voltage, for $V_{OUT} = \pm|V_{REF}$ ideal|. The polarity of the output voltage at this time will be opposite that of the applied reference. Finally, set all of the digital inputs HIGH and adjust "+full-scale adjust" for

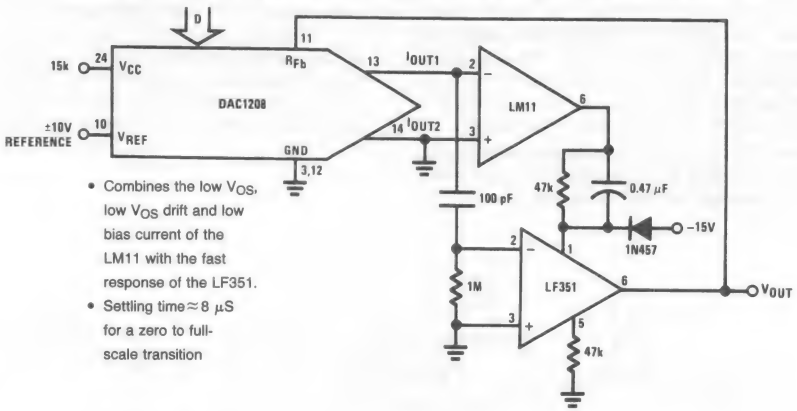
$$V_{OUT} = V_{REF}\frac{2047}{2048}.$$

The polarity of the output will be the same as that of the reference voltage.

3.0 APPLICATION IDEAS

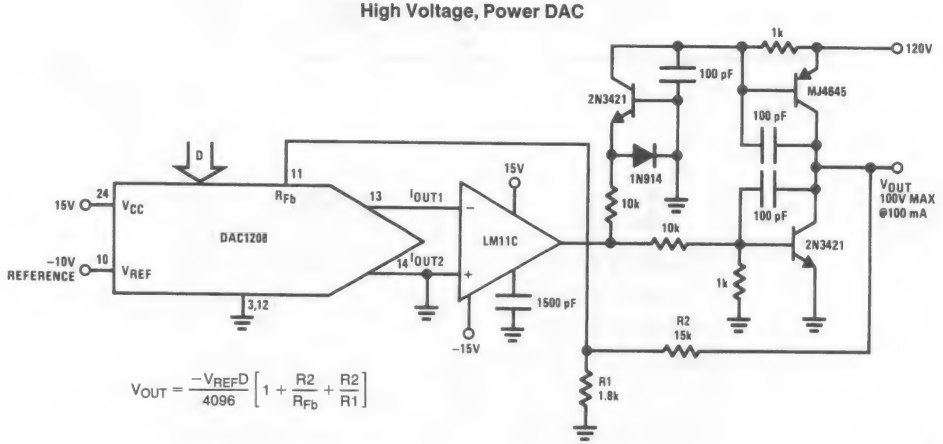
In this section the digital input word is represented by the letter D and is equal to the decimal equivalent of the 12-bit binary input. Hence D can be any integer value between 0 and 4095.

Composite Amplifier for Good DC Characteristics and Fast Output Response



- Combines the low V_{OS}, low V_{OS} drift and low bias current of the LM11 with the fast response of the LF351.
- Settling time≈8 μS for a zero to full-scale transition

High Voltage, Power DAC



$$V_{OUT} = \frac{-V_{REF}D}{4096}\left[1 + \frac{R2}{R_{Fb}} + \frac{R2}{R1}\right]$$

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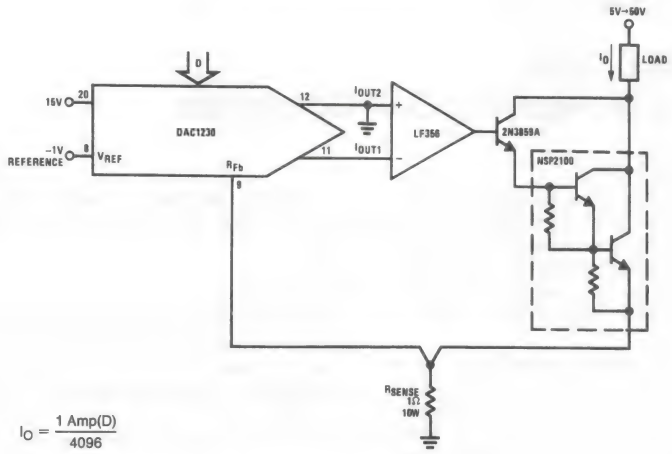
Application Hints (Continued)

High Current Controller



$$I_O = \frac{1\ \text{Amp}(D)}{4096}$$

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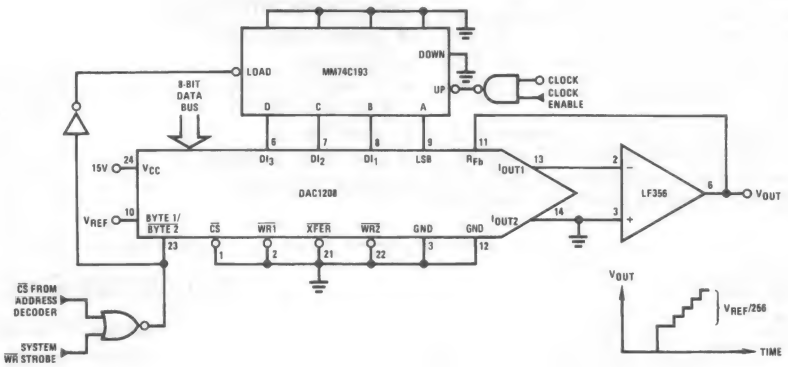
8-Bit Course, 4-Bit Vernier DAC



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Ordering Information

Part Number	Non-Linearity	Package	Temperature Range
DAC1208LCJ	0.012%	J24A Cerdip	−40°C to +85°C
DAC1208LCJ-1	0.012%	J24A Cerdip	0°C to +70°C
DAC1209LCJ	0.024%	J24A Cerdip	−40°C to +85°C
DAC1209LCJ-1	0.024%	J24A Cerdip	0°C to +70°C
DAC1210LCJ	0.050%	J24A Cerdip	−40°C to +85°C
DAC1210LCJ-1	0.050%	J24A Cerdip	0°C to +70°C
DAC1230LCJ	0.012%	J20A Cerdip	−40°C to +85°C
DAC1230LCJ-1	0.012%	J20A Cerdip	0°C to +70°C
DAC1231LCJ	0.024%	J20A Cerdip	−40°C to +85°C
DAC1231LCJ-1	0.024%	J20A Cerdip	0°C to +70°C
DAC1232LCJ	0.050%	J20A Cerdip	−40°C to +85°C
DAC1232LCJ-1	0.050%	J20A Cerdip	0°C to +70°C

National Semiconductor

DAC1218/DAC1219
12-Bit Binary Multiplying D/A Converter

General Description

The DAC1218 and the DAC1219 are 12-bit binary, 4-quadrant multiplying D to A converters. The linearity, differential non-linearity and monotonicity specifications for these converters are all guaranteed over temperature. In addition, these parameters are specified with standard zero and full-scale adjustment procedures as opposed to the impractical best fit straight line guarantee.

This level of precision is achieved though the use of an advanced silicon-chromium (SiCr) R-2R resistor ladder network. This type of thin-film resistor eliminates the parasitic diode problems associated with diffused resistors and allows the applied reference voltage to range from −25V to 25V, independent of the logic supply voltage.

CMOS current switches and drive circuitry are used to achieve low power consumption (20 mW typical) and minimize output leakage current errors (10 nA maximum). Unique digital input circuitry maintains TTL compatible input threshold voltages over the full operating supply voltage range.

The DAC1218 and DAC1219 are direct replacements for the AD7541 series, AD7521 series, and AD7531 series with a significant improvement in the linearity specification. In applications where direct interface of the D to A converter to a microprocessor bus is desirable, the DAC1208 and DAC1230 series eliminate the need for additional interface logic.

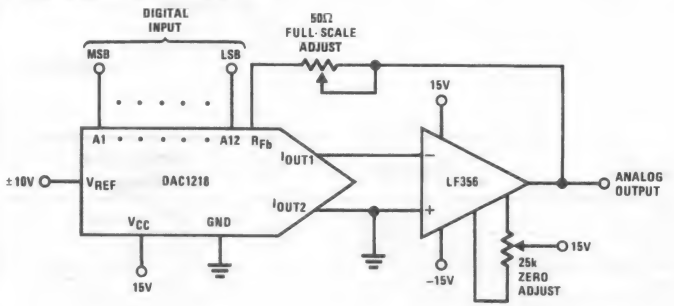
Features

- Linearity specified with zero and full-scale adjust only
- Logic inputs which meet TTL voltage level specs (1.4V logic threshold)
- Works with ±10V reference—full 4-quadrant multiplication
- All parts guaranteed 12-bit monotonic

Key Specifications

■ Current Settling Time	1 μs
■ Resolution	12 Bits
■ Linearity (Guaranteed over temperature)	12 Bits (DAC1218) 11 Bits (DAC1219)
■ Gain Tempco	1.5 ppm/°C
■ Low Power Dissipation	20 mW
■ Single Power Supply	5 V_{DC} to 15 V_{DC}

Typical Application



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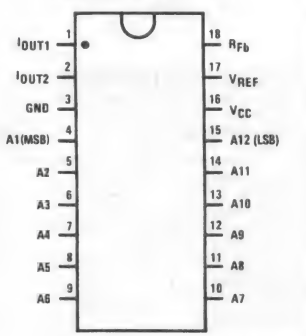
$$V_{OUT} = -V_{REF}\left(\frac{A1}{2} + \frac{A2}{4} + \frac{A3}{8} + \ldots \frac{A12}{4096}\right)$$

where: AN = 1 if digital input is high
AN = 0 if digital input is low

Connection Diagram

Dual-In-Line Package



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Top View

Ordering Information

Temperature Range		0°C to +70°C	−40°C to +85°C	Package Outline
Non Linearity	0.012%	DAC1218LCJ-1	DAC1218LCJ	J18A Cerdip
	0.024%	DAC1219LCJ-1	DAC1219LCJ	J18A Cerdip

Absolute Maximum Ratings (Notes 1 and 2)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_{CC})	17 V_{DC}
Voltage at Any Digital Input	V_{CC} to GND
Voltage at V_{REF} Input	±25V
Storage Temperature Range	−65°C to +150°C
Package Dissipation at $T_A = 25°C$ (Note 3)	500 mW
DC Voltage Applied to I_{OUT1} or I_{OUT2} (Note 4)	−100 mV to V_{CC}
Lead Temp. (Soldering, 10 seconds)	300°C
ESD Susceptibility (Note 11)	800V

Operating Conditions

Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
DAC1218LCJ, DAC1219LCJ	$-40°C \le T_A \le +85°C$
DAC1218LCJ-1, DAC1219LCJ-1	$0°C \le T_A \le 70°C$
Range of V_{CC}	5 V_{DC} to 16 V_{DC}
Voltage at Any Digital Input	V_{CC} to GND

Electrical Characteristics

V_{REF} = 10.000 V_{DC}, V_{CC} = 11.4 V_{DC} to 15.75 V_{DC} unless otherwise noted. **Boldface limits apply from T_{MIN} to T_{MAX} (see Note 9);** all other limits $T_A = T_J = 25°C$.

Parameter	Conditions	Notes	Typ (Note 10)	Tested Limit (Note 11)	Design Limit (Note 12)	Units
Resolution			12	**12**	**12**	Bits
Linearity Error (End Point Linearity)	Zero and Full-Scale Adjusted	4, 5, 9				
	DAC1218			**0.012**	**0.012**	% of FSR
	DAC1219			**0.024**	**0.024**	% of FSR
Differential Non-Linearity	Zero and Full-Scale Adjusted	4, 5, 9				
	DAC1218			**0.018**	**0.018**	% of FSR
	DAC1219			**0.024**	**0.024**	% of FSR
Monotonicity		4	12	12	**12**	Bits
Gain Error (Min)	Using Internal R_{Fb}, V_{REF} = ±10V, ±1V	5	−0.1	0.0		% of FSR
Gain Error (Max)		5	−0.1	−0.2		% of FSR
Gain Error Tempco		5	±1.3		±**6.0**	ppm of FS/°C
Power Supply Rejection	All Digital Inputs High	5	±3.0	±30		ppm of FSR/V
Reference Input Resistance	(Min)	9	15	**10**	**10**	kΩ
	(Max)	9	15	**20**	**20**	kΩ
Output Feedthrough Error	V_{REF} = 120 Vp-p, f = 100 kHz All Data Inputs Low	6	3.0			mVp-p
Output Capacitance	All Data Inputs High — I_{OUT1}				200	pF
	All Data Inputs High — I_{OUT2}				70	pF
	All Data Inputs Low — I_{OUT1}				70	pF
	All Data Inputs Low — I_{OUT2}				200	pF
Supply Current Drain		9		2.0	**2.5**	mA
Output Leakage Current		7, 9				
I_{OUT1}	All Data Inputs Low			**10**	**10**	nA
I_{OUT2}	All Data Inputs High			**10**	**10**	nA
Digital Input Threshold	Low Threshold	9		**0.8**	**0.8**	V_{DC}
	High Threshold			**2.2**	**2.2**	V_{DC}
Digital Input Currents	Digital Inputs <0.8V	9		−**200**	−**200**	μA_{DC}
	Digital Inputs >2.2V			**10**	**10**	μA_{DC}
t_s Current Settling Time	R_L = 100Ω, Output Settled to 0.01%, All Digital Inputs Switched Simultaneously		1			μs

Electrical Characteristics Notes

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: All voltages are measured with respect to GND, unless otherwise specified.

Note 3: This 500 mW specification applies for all packages. The low intrinsic power dissipation of this part (and the fact that there is no way to significantly modify the power dissipation) removes concern for heat sinking.

Note 4: Both I_{OUT1} and I_{OUT2} must go to ground or the virtual ground of an operational amplifier. The linearity error is degraded by approximately $V_{OS} \div V_{REF}$. For example, if $V_{REF} = 10V$ then a 1 mV offset, V_{OS}, on I_{OUT1} or I_{OUT2} will introduce an additional 0.01% linearity error.

Note 5: The unit FSR stands for full-scale range. Linearity Error and Power Supply Rejection specs are based on this unit to eliminate dependence on a particular V_{REF} value to indicate the true performance of the part. The Linearity Error specification of the DAC1218 is 0.012% of FSR. This guarantees that after performing a zero and full-scale adjustment, the plot of the 4096 analog voltage outputs will each be within 0.012% $\times$ V_{REF} of a straight line which passes through zero and full-scale. The unit ppm of FSR (parts per million of full-scale range) and ppm of FS (parts per million of full-scale) are used for convenience to define specs of very small percentage values, typical of higher accuracy converters. 1 ppm of FSR = $V_{REF}/10^6$ is the conversion factor to provide an actual output voltage quantity. For example, the gain error tempco spec of ± 6 ppm of FS/°C represents a worst-case full-scale gain error change with temperature from $-40°C$ to $+85°C$ of $\pm(6)(V_{REF}/10^6)(125°C)$ or ± 0.75 (10^{-3}) V_{REF} which is $\pm 0.075\%$ of V_{REF}.

Note 6: To achieve this low feedthrough in the D package, the user must ground the metal lid. If the lid is left floating the feedthrough is typically 6 mV.

Note 7: A 10 nA leakage current with $R_{Fb} = 20k$ and $V_{REF} = 10V$ corresponds to a zero error of $(10 \times 10^{-9} \times 20 \times 10^3) \times 100\%$ 10V or 0.002% of FS.

Note 8: Human body model, 100 pF discharged through 1.5 kΩ resistor.

Note 9: Tested limit for -1 suffix parts applies only at 25°C.

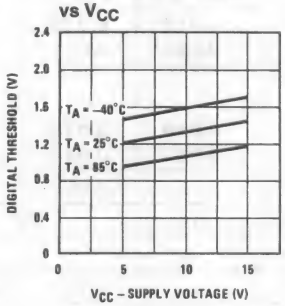
Note 10: Typicals are at 25°C and represent the most likely parametric norm.

Note 11: Tested limits are guaranteed to National's AOQL (Average Outgoing Quality Level).

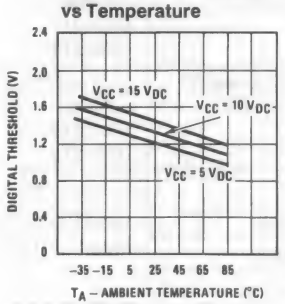
Note 12: Design limits are guaranteed but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Typical Performance Characteristics

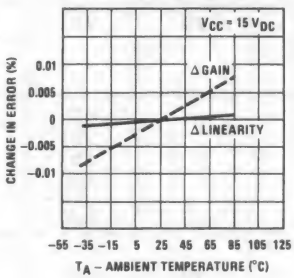
Digital Input Threshold vs V_{CC}



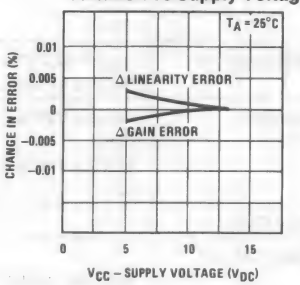
Digital Input Threshold vs Temperature



Gain and Linearity Error Variation vs Temperature



Gain and Linearity Error Variation vs Supply Voltage



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Definition of Package Pinouts

(A1–A12): Digital Inputs. A12 is the least significant digital input (LSB) and A1 is the most significant digital input (MSB).

I_{OUT1}: DAC Current Output 1. I_{OUT1} is a maximum for a digital input of all 1s, and is zero for a digital input of all 0s.

I_{OUT2}: DAC Current Output 2. I_{OUT2} is a constant minus I_{OUT1}, or $I_{OUT1} + I_{OUT2} =$ constant (for a fixed reference voltage).

R_{Fb}: Feedback Resistor. The feedback resistor is provided on the IC chip for use as the shunt feedback resistor for the external op amp which is used to provide an output voltage for the DAC. This on-chip resistor should always be used (not an external resistor) since it matches the resistors in the on-chip R-2R ladder and tracks these resistors over temperature.

V_{REF}: Reference Voltage Input. This input connects to an external precision voltage source to the internal R-2R ladder. V_{REF} can be selected over the range of 10V to −10V. This is also the analog voltage input for a 4-quadrant multiplying DAC application.

V_{CC}: Digital Supply Voltage. This is the power supply pin for the part. V_{CC} can be from 5 V_{DC} to 15 V_{DC}. Operation is optimum for 15 V_{DC}.

GND: Ground. This is the ground for the circuit.

Definition of Terms

Resolution: Resolution is defined as the reciprocal of the number of discrete steps in the DAC output. It is directly related to the number of switches or bits within the DAC. For example, the DAC1218 has 2^{12} or 4096 steps and therefore has 12-bit resolution.

Linearity Error: Linearity error in the maximum deviation from a *straight line passing through the endpoints of the DAC transfer characteristic.* It is measured after adjusting for zero and full scale. Linearity error is a parameter intrinsic to the device and cannot be externally adjusted.

National's linearity test (a) and the best straight line test (b) used by other suppliers are illustrated below. The best straight line (b) requires a special zero and FS adjustment for each part, which is almost impossible for the user to determine. The end point test uses a standard zero FS adjustment procedure and is a much more stringent test for DAC linearity.

Power Supply Sensitivity: Power supply sensitivity is a measure of the effect of power supply changes on the DAC full-scale output.

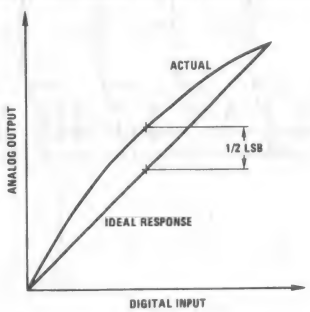
Settling Time: Full-scale current settling time requires zero to full-scale or full-scale to zero output change. Settling time is the time required from a code transition until the DAC output reaches within ± 1/2 LSB of the final output value.

Full-scale Error: Full-scale error is a measure of the output error between an ideal DAC and the actual device output. Ideally, for the DAC1218 full-scale is $V_{REF} - 1$ LSB. For $V_{REF} = 10V$ and unipolar operation, $V_{FULL\text{-}SCALE} = 10.0000V - 2.44\ mV = 9.9976V$. Full-scale error is adjustable to zero.

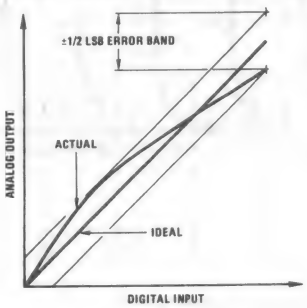
Differential Non-Linearity: The difference between any two consecutive codes in the transfer curve from the theoretical 1 LSB is differential non-linearity.

Monotonic: If the output of a DAC increases for increasing digital input code, then the DAC is monotonic. A 12-bit DAC which is monotonic to 12 bits simply means that input increasing digital input codes will produce an increasing analog output.

a) End point test after zero and FS adjust



b) Shifting FS adjust to pass best straight line test



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Application Hints

The DAC1218 and DAC1219 are pin-for-pin compatible with the DAC1220 series but feature 12 and 11-bit linearity specifications. To preserve this degree of accuracy, care must be taken in the selection and adjustments of the output amplifier and reference voltage. Careful PC board layout is important, with emphasis made on compactness of components to prevent inadvertent noise pickup and utilization of single point grounding and supply distribution.

1.0 BASIC CIRCUIT DESCRIPTION

Figure 1 illustrates the R-2R current switching ladder network used in the DAC1218 and DAC1219. As a function of the logic state of each digital input, the binarily weighted current in each leg of the ladder is switched to either I_{OUT1} or I_{OUT2}. The voltage potential at I_{OUT1} and I_{OUT2} must be at zero volts to keep the current in each leg the same, independent of the switch state.

The switches operate with a small voltage drop across them and can therefore conduct currents of either polarity. This permits the reference to be positive or negative, thereby allowing 4-quadrant multiplication by the digital input word. The reference can be a stable DC source or a bipolar AC signal within the range of $\pm 10V$, for specified accuracy, with an absolute maximum range of $\pm 25V$. The reference can also exceed the applied V_{CC} of the DAC.

The maximum output current from either I_{OUT1} or I_{OUT2} is equal to

$$\frac{V_{REF(max)}}{R}\left(\frac{4095}{4096}\right),$$

where R is the reference input resistance (typically 15 kΩ). A high level on any digital input steers current to I_{OUT1} and a low level steers current to I_{OUT2}.

2.0 CREATING A UNIPOLAR OUTPUT VOLTAGE (A DIGITAL ATTENUATOR)

To generate an output voltage and keep the potential at the current output terminals at 0V, an op amp current to voltage converter is used. As shown in *Figure 2*, the current from I_{OUT1} flows through the feedback resistor, forcing a proportional voltage at the amplifier output. The voltage at I_{OUT1} is held at a virtual ground potential. The feedback resistor is provided on the chip and should always be used as it matches and tracks the R value of the R-2R ladder. The output voltage is the opposite polarity of the applied reference voltage.

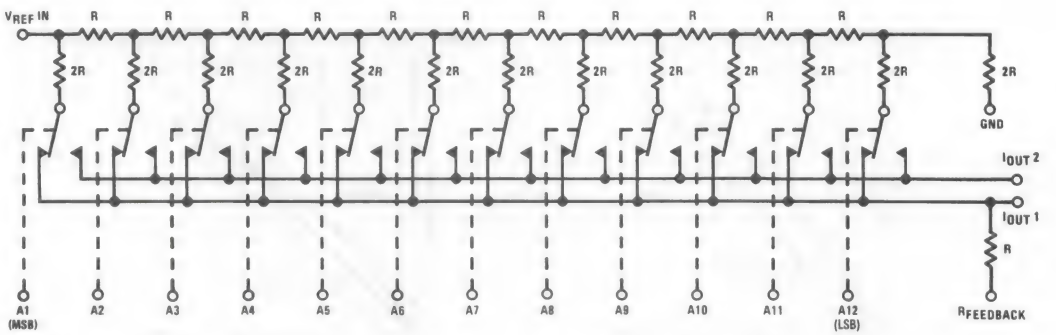
2.1 Amplifier Considerations

To maintain linearity of the output voltage with changing digital input codes the input offset voltage of the amplifier must be nulled. The resistance from I_{OUT1} to ground ($R_{I_{OUT1}}$) varies non-linearly with the applied digital code from a minimum of R with all ones applied to the input to near ∞ with an all zeros code. Any offset voltage between the amplifier inputs appears at the output with a gain of

$$1 + \frac{R_F}{R_{I_{OUT1}}}.$$

Since $R_{I_{OUT1}}$ varies with the input code, any offset will degrade output linearity. (See Note 4 of Electrical Characteristics.)

If the desired amplifier does not have offset balancing pins available (it could be part of a dual or quad package) the nulling circuit of *Figure 3* can be used. The voltage at the non-inverting input will be set to $-V_{OS}$ initially to force the inverting input to 0V. The common technique of summing current into the amplifier summing junction cannot be used as it directly introduces a zero code output current error.



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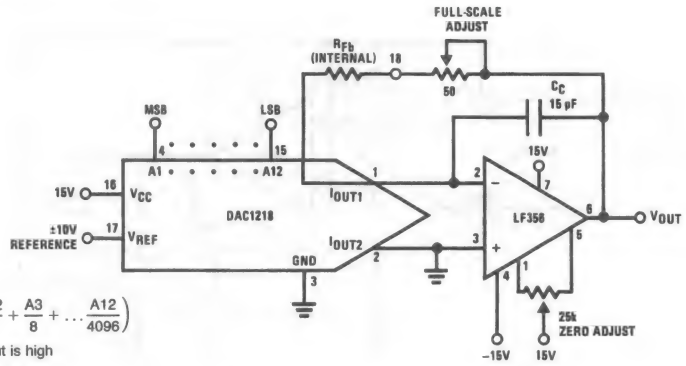
Note: Switches shown in digital high state.

FIGURE 1. The R-2R Current Switching Ladder Network

Application Hints (Continued)



$$V_{OUT} = -V_{REF}\left(\frac{A1}{2} + \frac{A2}{4} + \frac{A3}{8} + \ldots \frac{A12}{4096}\right)$$

where: AN=1 if digital input is high
AN=0 if digital input is low

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FIGURE 2. Unipolar Output Voltage

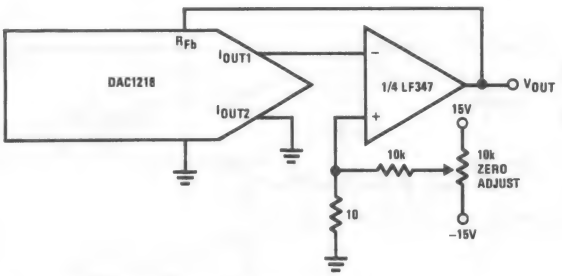

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FIGURE 3. Zeroing an Amplifier Which Does Not Have Balancing Provisions

The selected amplifier should have as low an input bias current as possible since input bias current contributes to the current flowing through the feedback resistor. BI-FET™ op amps such as the LF356 or LF351 or bipolar op amps with super β input transistors like the LM11 or LM308A produce negligible errors.

2.2 Zero and Full-Scale Adjustments

The fundamental purpose is to make the output voltages as near 0 V_{DC} as possible. This is accomplished in the circuit of *Figure 2* by shorting out the amplifier feedback resistance, and adjusting the V_{OS} nulling potentiometer of the op amp until the output reads zero volts. This is done, of course, with an applied digital input of all zeros if I_{OUT1} is driving the op amp (all ones for I_{OUT2}). The feedback short is then removed and the converter is zero adjusted.

A unique characteristic of these DACs is that any full-scale or gain error is always negative. This means that for a full-scale input code the output voltage, if not inherently correct, will always be less than what it should be. This ensures that adding an appropriate resistance in series with the internal feedback resistor, R_{Fb}, will always correct for any gain error. The 50Ω potentiometer in *Figure 2* is all that is needed to adjust the worst case DAC gain error.

Conversion accuracy is only as good as the applied reference voltage, so providing a source that is stable over time and temperature is important.

2.3 Output Settling Time

The output voltage settling time for this circuit in response to a change of the digital input code (a full-scale change is the worst case) is a combination of the DAC's output current settling characteristics and the settling characteristics of the output amplifier. The amplifier settling is further degraded by a feedback pole formed by the feedback resistance and the DAC output capacitance (which varies with the digital code). First order compensation for this pole is achieved by adding a feedback zero with capacitor C_C shown in *Figure 2*.

In many applications output response time and settling is just as important as accuracy. It can be difficult to find a single op amp that combines excellent DC characteristics (low V_{OS}, V_{OS} drift and bias current) with fast response and settling time. BI-FET op amps offer a reasonable compromise of high speed and good DC characteristics. The circuit of *Figure 4* illustrates a composite amplifier connection that combines the speed of a BI-FET LF351 with the excellent DC input characteristics of the LM11. If output settling time is not so critical, the LM11 can be used alone.

Figure 5 is a settling time test circuit for the complete voltage output DAC circuit. The circuit allows the settling time of the DAC amplifier to be measured to a resolution of 1 mV out of a zero to ±10V full-scale output change on an oscilloscope. *Figure 6* summarizes the measured settling times for several output amplifiers and feedback compensation capacitors.

Application Hints (Continued)

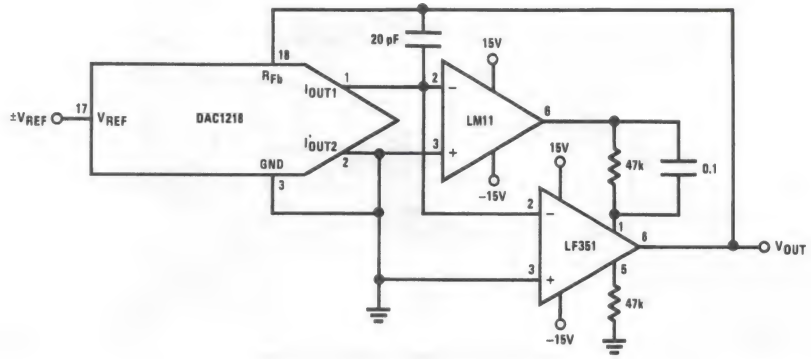

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FIGURE 4. Composite Output Amplifier Connection

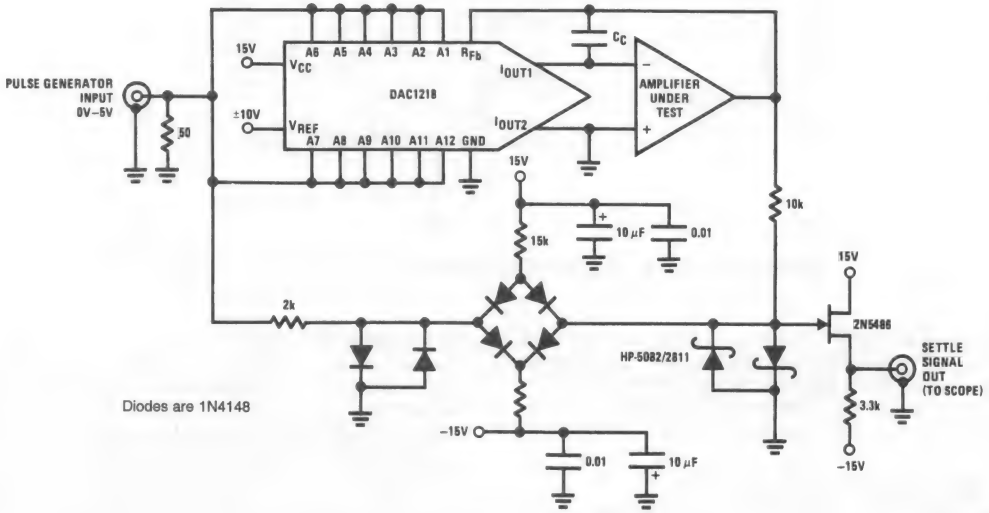

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FIGURE 5. DAC Settling Time Test Circuit

Amplifier	C_C	Settling Time to 0.01%
LM11	20 pF	30 μs
LF351	15 pF	8 μs
LF351	30 pF	5 μs
Composite LM11-LF351	20 pF	8 μs
LF356	15 pF	6 μs

FIGURE 6. Some Measured Settling Times

Application Hints (Continued)

3.0 OBTAINING A BIPOLAR OUTPUT VOLTAGE FROM A FIXED REFERENCE

The addition of a second op amp to the circuit of *Figure 2* can generate a bipolar output voltage from a fixed reference voltage (*Figure 7*). This, in effect gives sign significance to the MSB of the digital input word to allow two quadrant multiplication of the reference voltage. The polarity of the reference voltage can also be reversed to realize full 4-quadrant multiplication.

The output responds in accordance to the following expression:

$$V_O = V_{REF}\left(\frac{D - 2048}{2048}\right), 0 \le D \le 4095$$

where D is the decimal equivalent of the true binary input word. This configuration inherently accepts a code (half-scale or D = 2048) to provide 0V out without requiring an external ½ LSB offset as needed by other bipolar multiplying DAC circuits.

Only the offset voltage of amplifier A1 need be nulled to preserve linearity. The gain setting resistors around A2 must match and track each other. A thin film, 4-resistor network available from Beckman Instruments, Inc. (part no. 694-3-R10K-D) is ideally suited for this application. Two of the four resistors can be paralleled to form R and the other two can be used separately as the resistors labeled 2R.

Operation is summarized in the table below:

Applied Digital Input MSB											LSB	Decimal Equivalent	V_{OUT} $+V_{REF}$	V_{OUT} $-V_{REF}$
1	1	1	1	1	1	1	1	1	1	1	1	4095	$V_{REF} - 1$ LSB	$-\lvert V_{REF}\rvert + 1$ LSB
1	1	0	0	0	0	0	0	0	0	0	0	3072	$V_{REF}/2$	$-\lvert V_{REF}\rvert/2$
1	0	0	0	0	0	0	0	0	0	0	0	2048	0	0
0	1	1	1	1	1	1	1	1	1	1	1	2047	−1 LSB	+1 LSB
0	1	0	0	0	0	0	0	0	0	0	0	1024	$-V_{REF}/2$	$+\lvert V_{REF}\rvert/2$
0	0	0	0	0	0	0	0	0	0	0	0	0	$-V_{REF}$	$+\lvert V_{REF}\rvert$

Where 1 LSB = $\frac{|V_{REF}|}{2048}$

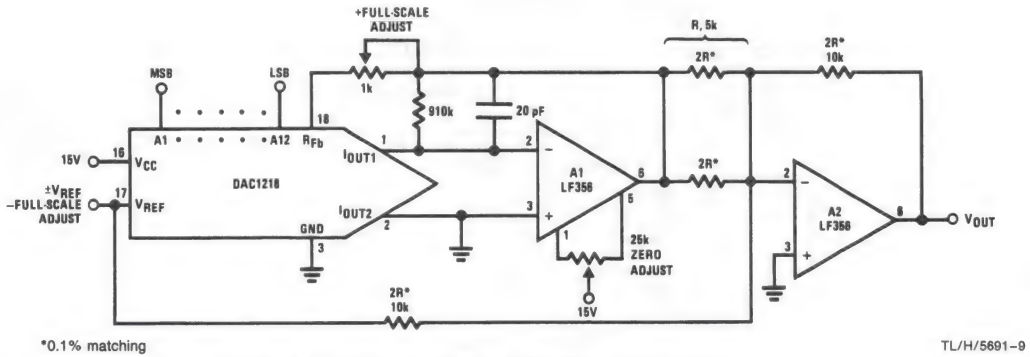


FIGURE 7. Obtaining a Bipolar Output from a Fixed Reference

Application Hints (Continued)

3.1 Zero and Full-Scale Adjustments

The three adjustments needed for this circuit are shown in *Figure 7*. The first step is to set all of the digital inputs LOW (to force I_{OUT1} to 0) and then trim "zero adjust" for zero volts at the inverting input (pin 2) of OA1. Next, with a code of all zeros still applied, adjust "- full-scale adjust", the reference voltage, for $V_{OUT} = \pm|(\text{ideal } V_{REF})|$. The sign of the output voltage will be opposite that of the applied reference.

Finally, set all of the digital inputs HIGH and adjust "+ full-scale adjust" for $V_{OUT} = V_{REF}$ (511/512). The sign of the output at this time will be the same as that of the reference voltage. This + full-scale adjustment scheme takes into account the effects of the V_{OS} of amplifier A2 (as long as this offset is less than 0.1% of V_{REF}) and any gain errors due to external resistor mismatch.

4.0 MISCELLANEOUS APPLICATION HINTS

The devices are CMOS products and reasonable care should be exercised in handling them to prevent catastrophic failures due to electrostatic discharge.

During power-up supply voltage sequencing, the negative supply of the output amplifier may appear first. This will typically cause the output of the op amp to bias near the negative supply potential. No harm is done to the DAC, however, as the on-chip 15 kΩ feedback resistor sufficiently limits the current flow from I_{OUT1} when this lead is clamped to one diode drop below ground.

As a general rule, any unused digital inputs should be tied high or low as required by the application. As a troubleshooting aid, if any digital input is left floating, the DAC will interpret that input as a logical 1 level.

Additional Application Ideas

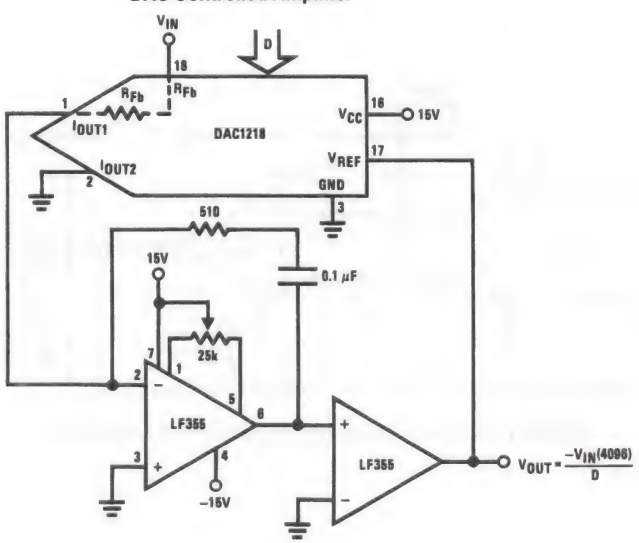
For the circuits shown, D represents the decimal equivalent of the binary digital input code. D ranges from 0 (for an all zeros input code) to 4095 (for an all ones input code) and for any code can be determined from:

$$D = 2048(A1) + 1024(A2) + 512(A2) + \ldots 2(A11) + 1(A12)$$

where AN = 1 if that input is high

AN = 0 if that input is low

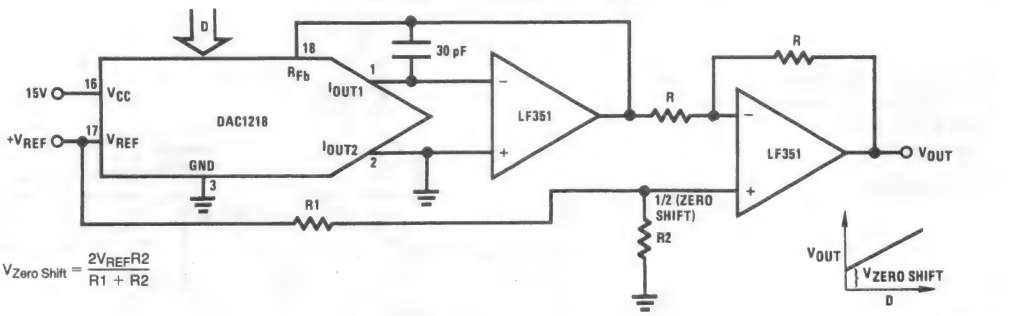
DAC Controlled Amplifier



TL/H/5691–10

Additional Application Ideas (Continued)

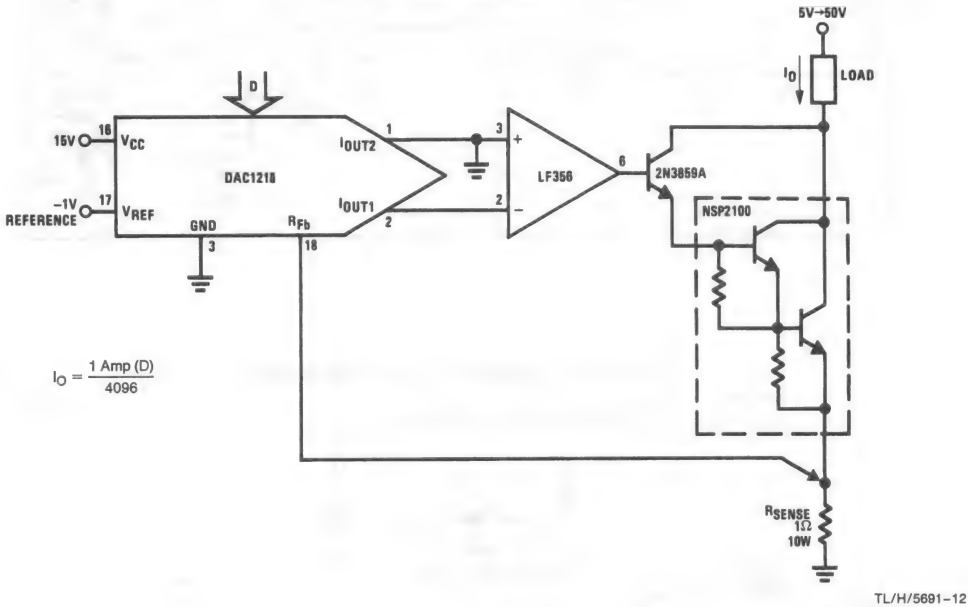
Offsetting the Zero Code Output Voltage



$$V_{Zero\ Shift} = \frac{2V_{REF}R2}{R1 + R2}$$

TL/H/5691–11

High Current Controller



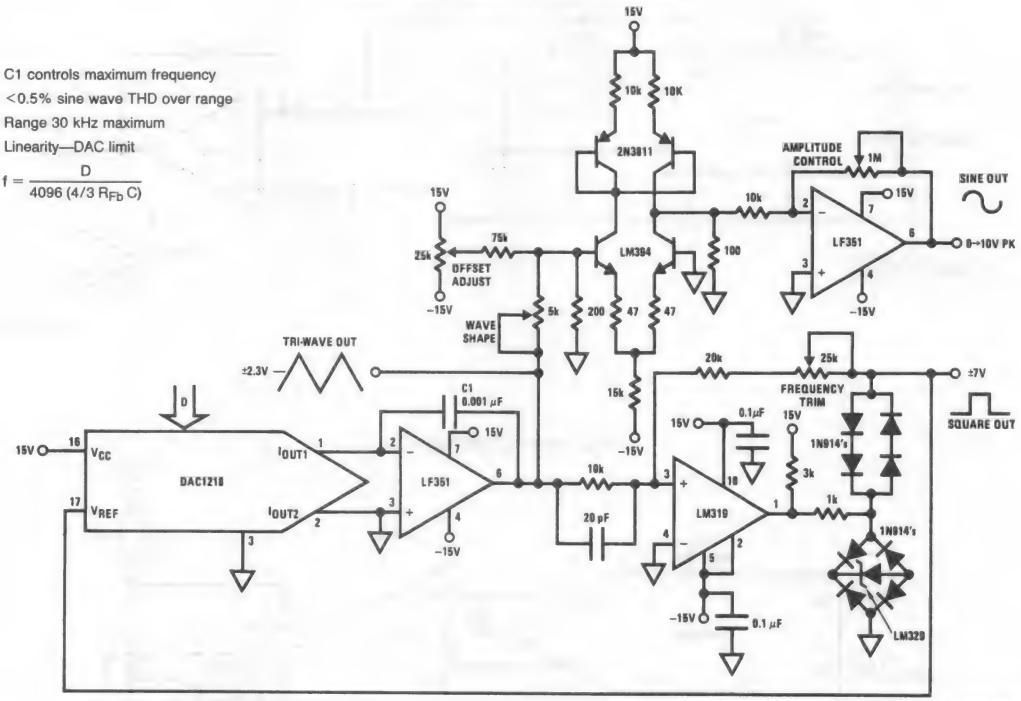
$$I_O = \frac{1\ Amp\ (D)}{4096}$$

TL/H/5691–12

Additional Application Ideas (Continued)

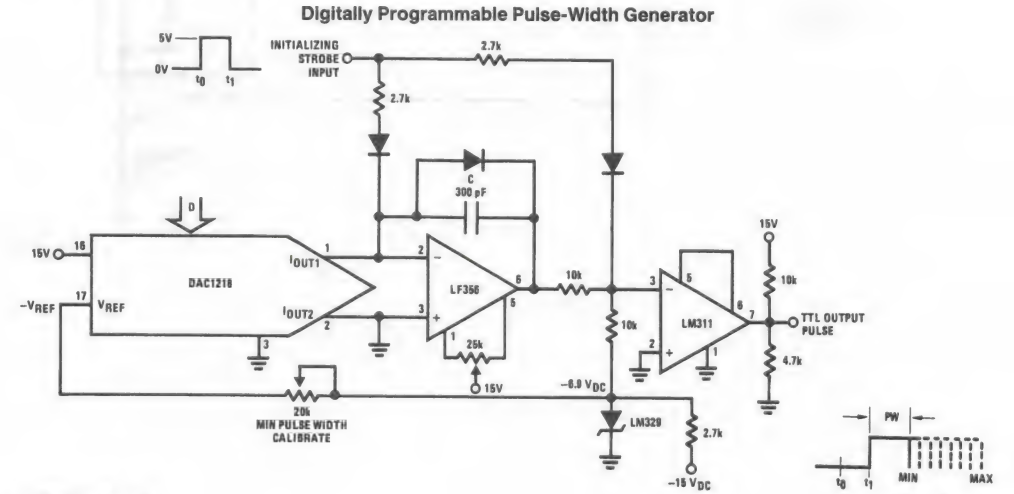
DAC Controlled Function Generator

- C1 controls maximum frequency
- <0.5% sine wave THD over range
- Range 30 kHz maximum
- Linearity—DAC limit
- $f = \frac{D}{4096\ (4/3\ R_{Fb}\ C)}$



TL/H/5691–13

Digitally Programmable Pulse-Width Generator



$$PW \cong \frac{C(7.5V)\ (4096)\ (R_{Fb})}{D|V_{REF}}$$

TL/H/5691–14

National Semiconductor

DAC1265A/DAC1265 Hi-Speed 12-Bit D/A Converter with Reference

General Description

The DAC1265A and DAC1265 are fast 12-bit digital to analog converters with internal voltage reference. These DACs use 12 precision high speed bipolar current steering switches, control amplifier, thin film resistor network, and buried zener voltage reference to obtain a high accuracy, very fast analog output current. The DAC1265A and DAC1265 have 10%–90% full-scale transition time under 35 ns and settle to less than ½ LSB in 200 ns. The buried zener reference has long-term stability and temperature drift characteristics comparable to the best discrete or separate IC references.

These digital to analog converters are recommended for applications in CRT displays, precision instruments and data acquisition systems requiring throughput rates as high as 5 MHz for full range transitions.

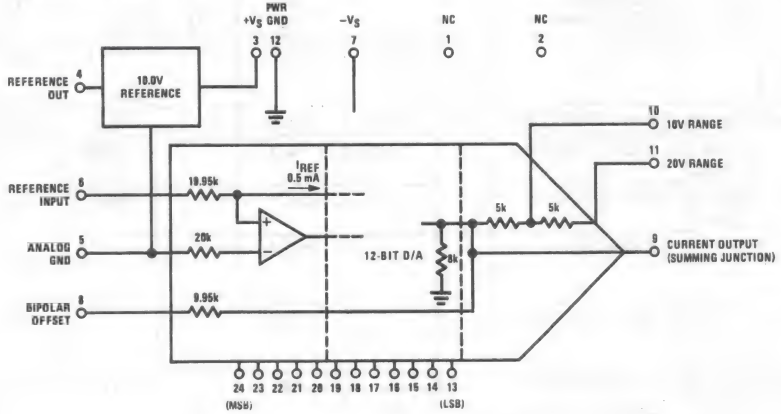
Features

- Bipolar current output DAC and voltage reference
- Fully differential, non-saturating precision current switch — R_{OUT} and C_{OUT} do not change with digital input code.
- Internal buried zener reference — 10V ± 1% max
- Precision thin film resistors for use with external op amp for voltage out or as input resistors for a successive approximation A/D converter
- Superior replacement for 12-bit D/A converters of this type

Key Specifications

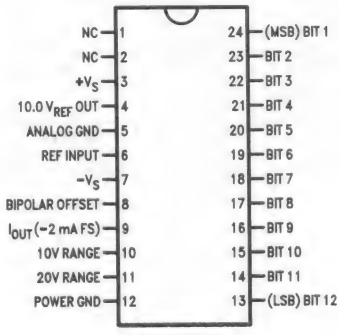
- Resolution and Monotonicity — 12 Bits
- Linearity — 12 Bits
  (Guaranteed over temperature)
- Output Current Settling Time — 400 ns max to 0.01%
- Gain Tempco — ±15 ppm/°C max
- Power Supply Sensitivity — ±10 ppm of FS/% V_{SUPPLY}

Block and Connection Diagrams



TL/H/5242–1

Dual-In-Line Package



Top View

TL/H/5242–2

Order Number DAC1265AJ, DAC1265ACJ, DAC1265LJ or DAC1265LCJ
See NS Package Number J24A

4

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V^+ and V^-)	±18V
Current Output (Pin 9) Voltage	−3V, 12V
Logic Input Voltage	−1V, 7V
Reference Input Voltage (Pin 6)	±12V
Analog GND to Power GND	±1V
Bipolar Offset	±12V
10V Range	±12V
20V Range	V^- to +24V
Power Dissipation (Note 1)	1000 mW
Short-Circuit Duration (Pins 4 to 12)	Continuous
Operating Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
DAC1265AJ, DAC1265LJ	−55°C to +125°C
DAC1265ACJ, DAC1265LCJ	0°C to +70°C
Storage Temperature Range	−65°C to +150°C
Maximum Junction Temperature	150°C
Lead Temperature (Soldering, 10 seconds)	300°C
ESD Susceptibility (Note 13)	TBD

Electrical Characteristics

$V_{SUPPLY} = \pm 15V \pm 5\%$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \le T_A \le T_{MAX}$.** For all other limits $T_A = 25°C$.

Parameter	Conditions		See Note	DAC1265A Typ (Note 11)	DAC1265A Tested Limit (Note 2)	DAC1265A Design Limit (Note 3)	DAC1265 Typ (Note 11)	DAC1265 Tested Limit (Note 2)	DAC1265 Design Limit (Note 3)	Units
CONVERTER CHARACTERISTICS										
Resolution					12			12		Bits
Linearity Error Max	Zero and Full-Scale Adjusted		4	±⅛	±¼ ±½		±¼	±½ ±¾		LSB
	AJ and LJ Suffix Parts ACJ and LCJ Suffix Parts					±½			±¾	
Differential Non-Linearity Max	Zero and Full-Scale Adjusted			±¼	±½		±½	±¾		
Monotonicity	AJ and LJ Suffix Parts ACJ and LCJ Suffix Parts				**12** 12	**12**		**12** 12	**12**	Bits
Full-Scale (Gain) Error Max	R2 = 50Ω in *Figure 1*		5	±0.1	±0.20		±0.1	±0.20		% Full-Scale
Offset Error Max All Bits OFF, Logic "0"	Unipolar (*Figure 1* Pin 8 Open)		6	±0.01	±0.05		±0.01	±0.05		
	Bipolar (R1 and R2 = 50Ω in *Figure 2*)		7	±0.05	±0.1		±0.05	±0.15		
Zero Error Max MSB ON	Bipolar (R1 and R2 = 50Ω in *Figure 2*)		8	±0.05	±0.1		±0.05	±0.15		
Gain Adjustment Range Min	R2 = 50Ω ± 50Ω in *Figure 1*				±0.2			±0.2		
Bipolar Offset Adjustment Range Min	R1 = 50Ω ± 50Ω and R2 = 50Ω in *Figure 2*				±0.15			±0.15		
Full-Scale (Gain) Temperature Coefficients Max	Using the Internal Reference	AJ and LJ Suffix	9	10	**15**		15	**30**		ppm/°C
		ACJ and LCJ Suffix		10		**20**	15		**50**	
Unipolar Offset Temperature Coefficients Max		AJ and LJ Suffix		1	**2**		1	**2**		
		ACJ and LCJ Suffix		1		**2**	1		**2**	
Bipolar Zero Temperature Coefficients Max		AJ and LJ Suffix		5	**10**		5	**10**		
		ACJ and LCJ Suffix		5		**10**	5		**10**	
Output Resistance	Exclusive of Offset and Range R_S			7.5	6 to 10		7.5	6 to 10		kΩ

Electrical Characteristics (Continued) $V_{SUPPLY} = \pm 15V \pm 5\%$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \le T_A \le T_{MAX}$.** For all other limits $T_A = 25°C$.

Parameter		Conditions	See Note	DAC1265A			DAC1265			Units
				Typ (Note 11)	Tested Limit (Note 2)	Design Limit (Note 3)	Typ (Note 11)	Tested Limit (Note 2)	Design Limit (Note 3)	
Current Output		Unipolar		−2	−1.6 to −2.4		−2	−1.6 to −2.4		mA
		Bipolar		±1.0	±0.8 to ±1.2		±1.0	±0.8 to ±1.2		
Output Capacitance				25			25			pF
Output Noise (FS, 10V Range)		10 Hz to 100 kHz with Internal Reference		40			40			μVrms
Typ Output Voltage Ranges		Using Internal Offset and Range R_S		±2.5, ±5, ±10, 0 to 5, 0 to 10						V
Reference Input Resistance				20.8	15 to 25		20.8	15 to 25		kΩ
Output Compliance Voltage						**−1.5 to 10**			**−1.5 to 10**	V
REFERENCE OUTPUT CHARACTERISTICS										
Reference Voltage	Min	$I_{REF} = 1.5$ mA		10.00	9.90		10.00	9.90		V
	Max				10.10			10.10		
Temperature Coefficient				±8			±12			ppm/°C
Reference Output Current Min					3.0			3.0		mA
Output Resistance Max		$f_O = 1$ kHz, 0.5 mA ≤ I_{REF} ≤ 3 mA		0.05	1.0		0.05	1.0		Ω
DIGITAL AND DC CHARACTERISTICS										
Logic Input Voltage		Logic High Bit ON — AJ and LJ Suffix			**2 to 5.5**			**2 to 5.5**		V
		Logic High Bit ON — ACJ and LCJ Suffix			1.9 to 5.5	**2 to 5.5**		1.9 to 5.5	**2 to 5.5**	
	Max	Logic Low Bit OFF — AJ and LJ Suffix			**0.8**			**0.8**		
		Logic Low Bit OFF — ACJ and LCJ Suffix			1.0	**0.8**		1.0	**0.8**	
Logic Input Current Max		Logic High — AJ and LJ Suffix		150	**300**		150	**300**		μA
		Logic High — ACJ and LCJ Suffix		150	280	**300**	150	280	**300**	
		Logic Low — AJ and LJ Suffix		45	**100**		45	**100**		
		Logic Low — ACJ and LCJ Suffix		45	90	**100**	45	90	**100**	
Power Supply Current Max	I+	V^+ Supply = 15V ± 10%		3	5		3	5		mA
	I−	V^- Supply = −15V ± 10%		−12	−18		−12	−18		
Power Dissipation Max		$V_{SUPPLY} = \pm 15V$		225	345		225	345		mW
Power Supply Sensitivity Max		V^+ Supply = 15V ± 10%	10	±3	±10		±3	±10		ppm of FS/% V_{SUPPLY}
		V^- Supply = −15V ± 10%	10	±15	±25		±15	±25		

Electrical Characteristics (Continued) $V_{SUPPLY} = \pm 15V \pm 5\%$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \leq T_A \leq T_{MAX}$.** For all other limits $T_A = 25°C$.

Parameter	Conditions	See Note	DAC1265A			DAC1265			Units
			Typ (Note 11)	Tested Limit (Note 2)	Design Limit (Note 3)	Typ (Note 11)	Tested Limit (Note 2)	Design Limit (Note 3)	
AC CHARACTERISTICS									
Settling Time Max	FSR Change		200		400	200		400	ns
Full-Scale Transition Max	10% to 90% Rise Time Plus Delay Time		15		30	15		30	ns
	90% to 10% Fall Time Plus Delay Time		30		50	30		50	

Note 1: The typical θ_{JA} of the 24-pin package is 80° C/W.

Note 2: Tested and guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 3: Guaranteed, but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Note 4: Linearity error $= \frac{V_{OUT} - V_{OFFSET} - (D \times V_{LSB})}{V_{LSB}}$ where $V_{LSB} = \frac{V_{FS} - V_{OFFSET}}{4095}$ and D is the digital input (0 to 4095) which produced V_{OUT}.

Note 5: Percent gain error for 10V range $= \frac{(V_{FS} - V_{OFFSET}) - (4095/4096)10V}{10V} \times 100$.

Note 6: Unipolar offset error for 10V range $= (V_{OUT}/10V) \times 100$ in percent of full-scale.

Note 7: Bipolar offset error for 10V range $= \frac{V_{OUT} - (-5V)}{10V} \times 100$ in percent of full-scale.

Note 8: Bipolar zero error for 10V range $= (V_{OUT}/10V) \times 100$ in percent of full-scale.

Note 9: Gain error tempco $= \frac{(V_{FS} - V_{OFFSET}) \text{at } (T_{MAX} \text{ or } T_{MIN}) - (V_{FS} - V_{OFFSET}) \text{ at } 25°C}{10V \text{ range} \times (T_{MAX} \text{ or } T_{MIN} - 25°C)} \times 10^6$ in ppm/°C.

Note 10: Power supply sensitivity for 10V range $= 10^6 \times \frac{(V_{FS} - V_{OFFSET}) \text{ at } (16.5V \text{ or } -13.5V) - (V_{FS} - V_{OFFSET}) \text{ at } (13.5V \text{ or } -16.5V)}{10V \times 20\%}$ in ppm of FS/%V_S.

The opposite supply is held at −15V or +15V respectively.

Note 11: Typicals are at 25°C and represent most likely parametric norm.

Note 12: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 13: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Functional Description and Applications

1.0 BUFFERED VOLTAGE OUTPUT CONNECTION

The standard current-to-voltage conversion connections using an operational amplifier are shown here with the preferred trimming techniques. If a low offset operational amplifier (LF401A) is used, excellent performance can be obtained in many situations without trimming (an op amp with less than 0.5 mV maximum offset voltage should be used to keep offset errors below ½ LSB). Unipolar zero will typically be within ±½ LSB (plus op amp offset), and if a 50Ω fixed resistor is substituted for the 100Ω trimmer (R2, *Figure 1*), full-scale accuracy will be within 0.1% (0.20% maximum). Substituting a 50Ω resistor for the 100Ω bipolar offset trimmer (R1, *Figure 2*) will give a bipolar zero error typically within ±2 LSB (0.05%).

1.1 Unipolar Configuration *(Figure 1)*

This configuration will provide a unipolar 0V to 9.9976V output range.

Step 1—Offset Adjust (Zero)

Turn all bits OFF and adjust zero trimmer, R1, until the output reads 0.000V (1 LSB = 2.44 mV). In most cases this trim is not needed.

Step 2—Gain Adjust

Turn all bits ON and adjust 100Ω gain trimmer, R2, until the output is 9.9976V (full-scale adjusted to 1 LSB less than nominal full-scale of 10.000V). If a 10.2375V full-scale is desired (exactly 2.5 mV/bit), insert a 120Ω resistor in series with the gain resistor at pin 10 to the op amp output.

1.2 Bipolar Configuration *(Figure 2)*

This configuration will provide a bipolar output voltage from — 5.000V to 4.9976V, with positive full-scale occurring with all bits ON (all 1s).

Step 1—Offset Adjust

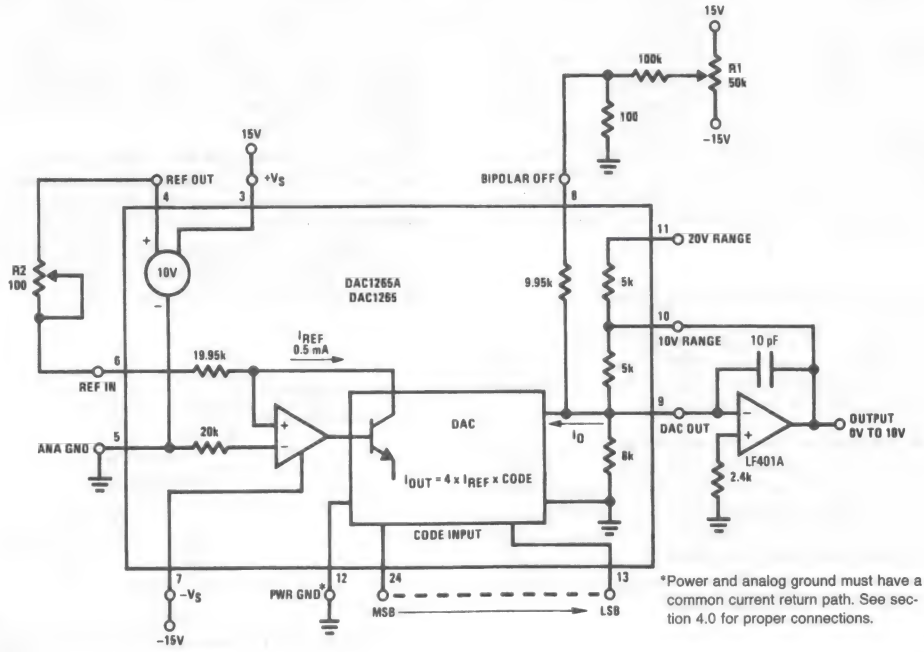
Turn OFF all bits. Adjust 100Ω offset trimmer, R1, to give −5.000V output.

Step 2—Gain Adjust

Turn ON all bits. Adjust 100Ω gain trimmer, R2, to give a reading of 4.9976V.

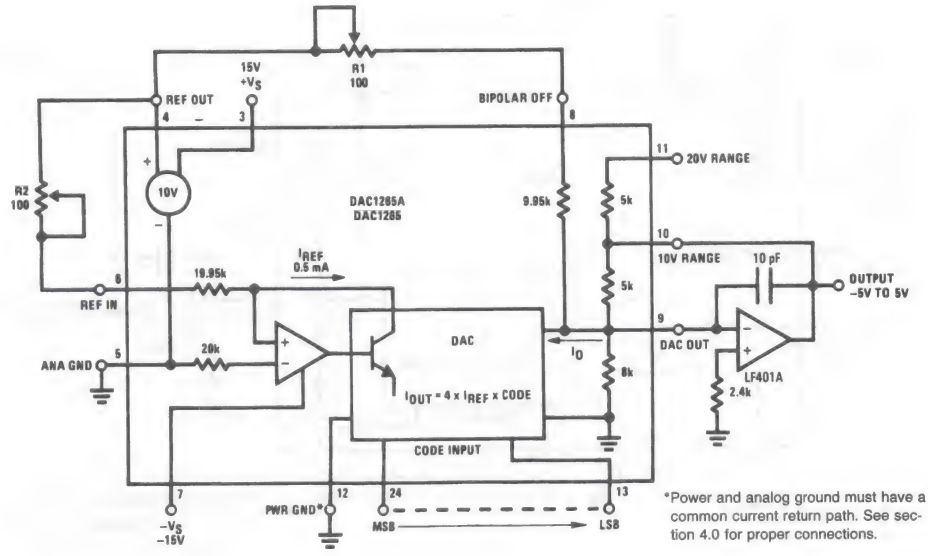
Please note that it is not necessary to trim the op amp to obtain full accuracy at room temperature. In most bipolar situations, an op amp trim is unnecessary unless the untrimmed offset drift of the op amp is excessive. Bipolar zero error (MSB bit ON) is not adjusted separately and is typically < ±0.05% of FS after offset and gain adjust.

Functional Description and Applications (Continued)



TL/H/5242–3

FIGURE 1. 0V to 10V Unipolar Voltage Output



TL/H/5242–5

FIGURE 2. ±5V Bipolar Voltage Output

Functional Description and Applications (Continued)

1.3 Other Voltage Ranges *(Figure 3)*

The DAC1265A and DAC1265 can also be easily configured for a unipolar 0V to 5V range or ±2.5V and ±10V bipolar ranges by using the additional 5k application resistor provided at the 20V range R terminal, pin 11. For a 5V range (0V to 5V or ±2.5V), the two 5k resistors are used in parallel by shorting pin 11 to pin 9 and connecting pin 10 to the op amp output and the bipolar offset either left open for unipolar or connected through a 100Ω pot to the REF OUT for the bipolar range. For the ± 10V range use the 5k resistors in series by connecting only pin 11 to the op amp output and connecting the bipolar offset as shown. The ±10V option is shown in *Figure 3.*

2.0 INTERNAL/EXTERNAL REFERENCE USE

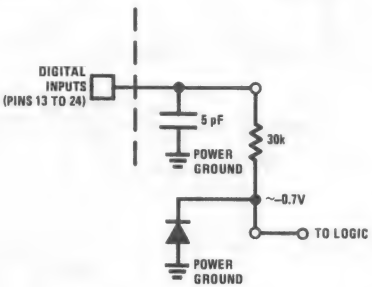
The performance of the DAC1265A and DAC1265 is specified with the internal reference driving the DAC since all trimming and testing (especially for full-scale error and bipolar operation) is done in this configuration.

The internal reference has sufficient buffering to drive external circuitry in addition to the reference currents required for the DAC (typically 0.5 mA to REF IN and 1.0 mA to BIPOLAR OFFSET, if used). A minimum of 1.5 mA is available for driving external circuits. The reference is typically trimmed to ±0.2%, then tested and guaranteed to ±1.0% maximum error. The temperature coefficient is comparable to that of the full-scale TC for a particular grade.

3.0 DIGITAL INPUT

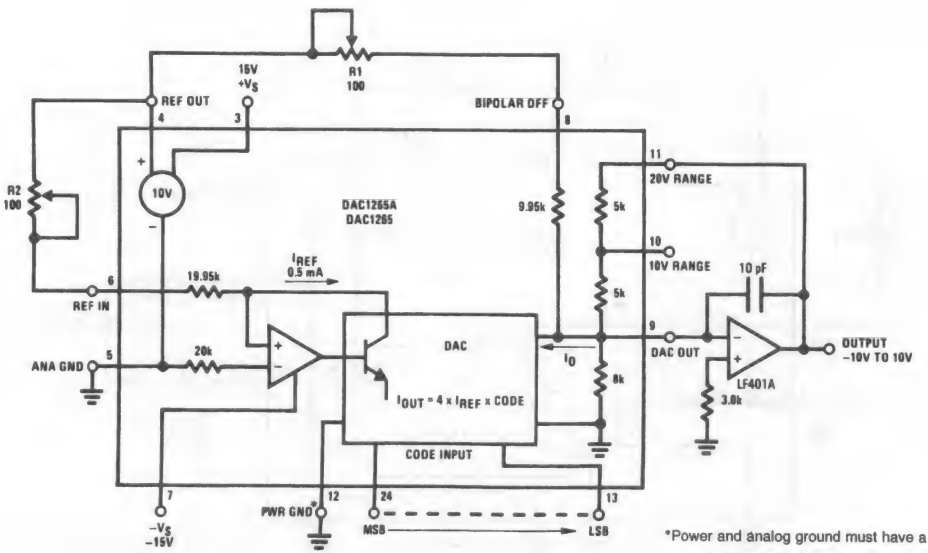
The DAC1265A and DAC1265 use a standard positive true straight binary code for unipolar outputs (all 1s give full-scale output), and an offset binary code for bipolar output ranges. In the bipolar mode, with all 0s on the inputs, the output will go to negative full-scale; with 100...00 (only the MSB on), the output will be 0.00V; with all 1s, the output will go to positive full-scale.

The threshold of the digital input circuitry is set at 1.4V and does not vary with supply voltage. The input lines can interface with any type of 5V logic, TTL/DTL or CMOS, and have sufficiently low input currents to interface easily with unbuffered CMOS logic. The configuration of the input circuit is shown in *Figure 4.* The input line can be modeled as a 30 kΩ resistance connected to a −0.7V rail.



TL/H/5242–6

FIGURE 4. Equivalent Digital Input Circuit



*Power and analog ground must have a common current return path. See section 4.0 for proper connections.

TL/H/5242–4

FIGURE 3. ±10V Voltage Output

Functional Description and Applications (Continued)

4.0 APPLICATION OF ANALOG AND POWER GROUNDS

The DAC1265A and DAC1265 have separate analog and power ground pins to allow optimum connections for low noise and high speed performance. The two ground lines can be separated by up to 200 mV without any loss in performance. There may be some loss in linearity beyond that level. If these DACs are to be used in a system in which the two grounds will be ultimately connected at some distance from the device, it is recommended that parallel back-to-back diodes be connected between the ground lines near the device to prevent a fault condition.

The analog ground at pin 5 is the ground reference point for the internal reference and is thus the "high quality" ground; it should be connected directly to the analog reference point of the system. The power ground at pin 12 can be connected to the most convenient ground reference point; analog power return is preferred, but digital ground is acceptable. If power ground contains high frequency noise beyond 200 mV, this noise may feed through the converter, so that some caution will be required in applying these grounds.

5.0 OUTPUT VOLTAGE COMPLIANCE

The DAC1265A and DAC1265 have a typical output compliance range from −2V to 10V. The current-steering output stages will be unaffected by changes in the output terminal voltage over that range. However, there is an equivalent output impedance of 8k in parallel with 25 pF at the output terminal which produces an equivalent error current if the voltage deviates from power ground. This is a linear effect that does not change with input code. Operation beyond the compliance limits may cause either output stage saturation or breakdown which results in non-linear performance. Compliance limits are not affected by the positive power supply, but are a function of output current and negative supply.

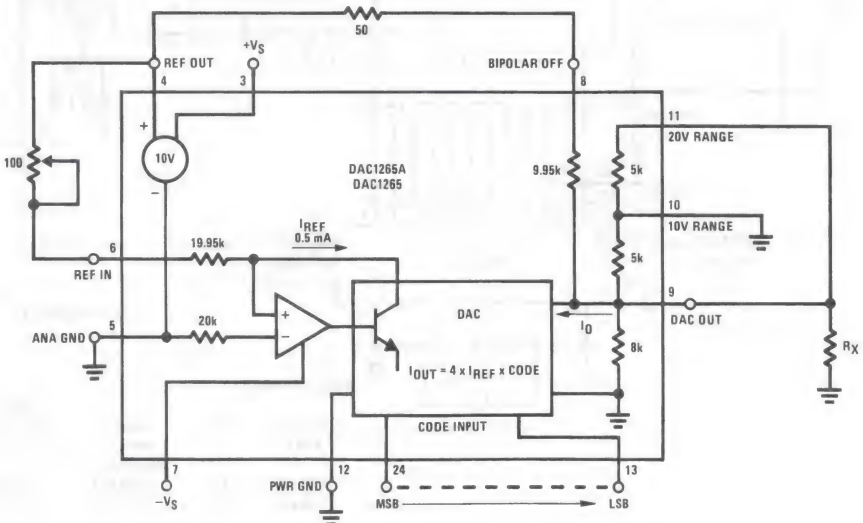
6.0 DIRECT UNBUFFERED VOLTAGE OUTPUT FOR CABLE DRIVING

The wide compliance range allows direct current-to-voltage conversion with just an output resistor. *Figure 5* shows a connection using the gain and bipolar output resistors to give a ±1.60V bipolar swing. In this situation, the digital code is complementary binary. Other combinations of internal and external output resistors (R_X) can be used to scale to alternate voltage ranges, simply by appropriately scaling the 0 mA to −2 mA unipolar output current and using the 10.0V reference voltage for bipolar offset. For example, setting $R_X = 2.67\ k\Omega$ gives a ±1V range with a 1 kΩ equivalent output impedance.

This connection is especially useful for directly driving a long cable at high speed. Using a 50Ω resistor for R_X would allow interface to a 50Ω cable with a ±50 mV full-scale swing.

7.0 HIGH SPEED 12-BIT A/D CONVERTERS

The fast settling characteristics of the DAC1265A and DAC1265 make them ideal for high speed successive approximation A/D converters. The internal reference and trimmed internal resistors allow a 12-bit converter system to be constructed with a minimum parts count. Shown in *Figure 6* is a configuration using standard components; this system completes a full 12-bit conversion in 10 μs unipolar or bipolar. This converter will be accurate to ±½ LSB of 12 bits and have a typical gain TC of 10 ppm/°C.



TL/H/5242–7

FIGURE 5. Unbuffered Bipolar Voltage Output

Functional Description and Applications (Continued)

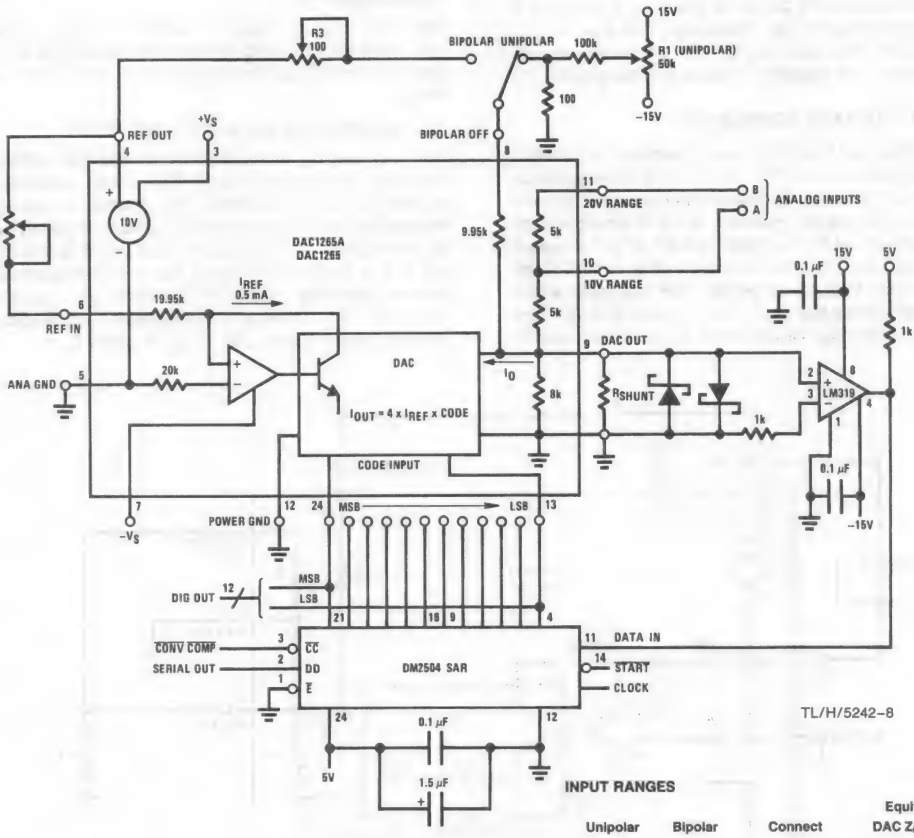
In the unipolar mode, the system range is 0V to 9.9976V, with each bit having a value of 2.44 mV. For true conversion accuracy, an A/D converter should be trimmed so that a given output code results from input levels from ½ LSB below to ½ LSB above the exact voltage represented by that code. Therefore, the converter zero point should be trimmed with an input voltage of 1.22 mV; trim R1 until the LSB just begins to appear in the output code (all other bits "0"). For full-scale, use an input voltage of 9.9963V (10V-1 LSB-½ LSB); then trim R2 until the LSB just begins to appear (all other bits "1").

The bipolar signal range is −5.0V to 4.9976V. Bipolar offset trimming is done by applying a −4.9988V input signal and trimming R3 for the LSB transition (all other bits "0").

Full-scale is set by applying 4.9963V and trimming R2 for the LSB transition (all other bits "1"). In many applications, the pretrimmed internal resistors are sufficiently accurate that external trimmers will be unnecessary, especially in situations requiring less than full 12-bit ±½ LSB accuracy.

For fastest operation, the impedance at the comparator summing node must be minimized. However, lowering the impedance will reduce the voltage signal to the comparator (at an equivalent impedance at the summing node of 1 kΩ, 1 LSB = 0.5 mV), to the point that comparator performance will be sacrificed. The contribution to this impedance from the DAC will vary with the input configuration (*Figure 6*, Input Ranges Table).

To prevent dynamic errors, the input signal should have a low dynamic source impedance, such as that of the LF411A op amp.



INPUT RANGES

Unipolar	Bipolar	Connect	Equiv. DAC Z_{OUT}
0 to 5	±2.5	Input to A B to DAC OUT	1.60 kΩ
0 to 10	±5	Input to A	2.35 kΩ
0 to 20	±10	Input to B	3.08 kΩ

FIGURE 6. Fast Precision Analog to Digital Converter

Definition of Terms

Digital Inputs: The DAC1265A and DAC1265 accept digital input codes in binary format and may be user connected for any one of three binary codes: straight binary, two's complement, or offset binary.

Digital Input MSB LSB	Analog Output		
	Straight Binary	Offset Binary	Two's Complement*
000..000	zero	−FS (Full-Scale)	zero
011...111	½ FS−1 LSB	zero−1 LSB	+FS−1 LSB
100..000	½ FS	zero	−FS
111...111	+FS−1 LSB	+FS−1 LSB	zero−1 LSB

*Invert MSB with external inverter to obtain Two's Complement coding

Linearity Error: Linearity error of a D/A converter is an important measure of its accuracy. It describes the deviation from an ideal straight line transfer curve drawn between zero (all bits OFF) and full-scale (all bits ON).

Differential Non-Linearity: For a D/A converter, it is the difference between the actual output voltage change and the ideal (1 LSB) voltage change for a one-bit change in code. A differential non-linearity of ±1 LSB or less guarantees monotonicity; i.e., the output always increases and never decreases for an increasing input. It is guaranteed by testing the major carry transitions, i.e., 100...000 to 011...111, etc.

Settling Time: Settling time is the time required for the output to settle to within the specified error band for any input code transition. It is usually specified for a full-scale or major carry transition.

Gain Tempco: The change in full-scale analog output over the specified temperature range expressed in parts per million of full-scale per °C (ppm of FS/°C). Gain error is measured with respect to 25°C at high (T_{MAX}) and low (T_{MIN}) temperatures. Gain tempco is calculated for both high ($T_{MAX}-25°C$) and low ($25°C-T_{MIN}$) ranges by dividing the gain error by the respective change in temperature. The specification is the larger of the two representing worst-case drift.

Offset Tempco: The change in analog output with all bits OFF over the specified temperature range expressed in parts per million of full-scale per °C (ppm of FS/°C). Offset error is measured with respect to 25°C at high (T_{MAX}) and low (T_{MIN}) temperatures. Offset tempco is calculated for both high ($T_{MAX}-25°C$) and low ($25°C-T_{MIN}$) ranges by dividing the offset error by the respective change in temperature. The specification given is the larger of the two, representing worst-case drift.

Power Supply Sensitivity: Power supply sensitivity is a measure of the change in gain and offset of the D/A converter resulting from a change in −15V or + 15V supplies. It is specified under DC conditions and expressed as parts per million of full-scale per percent of change in power supply (ppm of FS/%).

Ordering Information

Temperature Range		0°C to 70°C	−55°C to + 125°C
Linearity Error Over Temperature	±½ Bit	DAC1265ACJ	DAC1265AJ
	±¾ Bit	DAC1265LCJ	DAC1265LJ

4

National Semiconductor

DAC1266A/DAC1266 Hi-Speed 12-Bit D/A Converter

General Description

The DAC1266A and DAC1266 are fast 12-bit digital to analog converters. These DACs use 12 precision high speed bipolar current steering switches, control amplifier, and a thin film resistor network to obtain a high accuracy, very fast analog output current. The DAC1266A and DAC1266 have 10%–90% full-scale transition time under 30 ns and settle to less than ½ LSB in 200 ns.

These digital to analog converters are recommended for applications in CRT displays, precision instruments and data acquisition systems requiring throughput rates as high as 5 MHz for full range transitions.

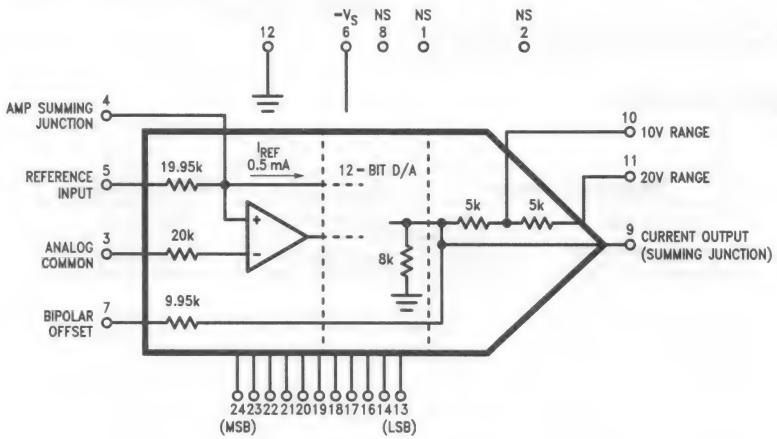
Features

- Bipolar current output DAC
- Fully differential, non-saturating precision current switch — R_{OUT} and C_{OUT} do not change with digital input code
- Precision thin film resistors for use with external op amp for voltage out or as input resistors for a successive approximate A/D converter
- Superior replacement for 12-bit D/A converters of this type

Key Specifications

- Resolution and Monotonicity 12 Bits
- Linearity 12 Bits
  (Guaranteed over temperature)
- Output Current Settling Time 400 ns max to 0.01%
- Full-Scale Transition Time (10%–90%) 30 ns
- Power Supply Sensitivity ±15 ppm of FS/% V_{SUPPLY}

Block and Connection Diagrams



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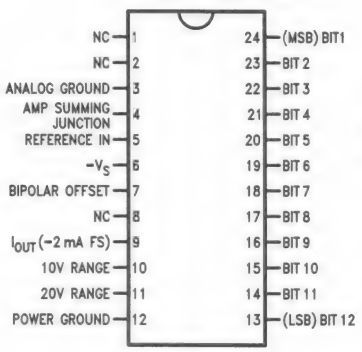
Dual-In-Line Package



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Top View

Order Number
DAC1266AJ, DAC1266ACJ,
DAC1266LJ or DAC1266LCJ
See NS Package Number J24A

Absolute Maximum Ratings (Note 11)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V^-)	0V to −18V
Current Output (Pin 9) Voltage	−3V, 12V
Logic Input Voltage	−1V, 7V
Reference Input Voltage (Pin 5)	±12V
Analog GND to Power GND	±1V
Bipolar Offset	±12V
10V Range	±12V
20V Range	V^- to +24V
Power Dissipation (Note 1)	1000 mW
Operating Temperature Range	$T_{MIN} \le T_A \le T_{MAX}$
DAC1266AJ, DAC1266LJ	−55°C to +125°C
DAC1266ACJ, DAC1266LCJ	0°C to +70°C
Storage Temperature Range	−65°C to +150°C
Maximum Junction Temperature	150°C
Lead Temp. (Soldering, 10 sec.)	300°C
ESD Susceptibility (Note 12)	TBD

Electrical Characteristics

$V_{SUPPLY} = -15V \pm 5\%$ and $V_{REF} = 10.000V$ unless otherwise noted. **Boldface limits apply over temperature, $T_{MIN} \le T_A \le T_{MAX}$.** For all other limits $T_A = 25°C$.

Parameter	Conditions	See Note	DAC1266A			DAC1266			Units
			Typ	Tested Limit (Note 2)	Design Limit (Note 3)	Typ	Tested Limit (Note 2)	Design Limit (Note 3)	
CONVERTER CHARACTERISTICS									
Resolution				12			12		Bits
Linearity Error Max	Zero and Full-Scale Adjusted	4	±⅛	±¼		±¼	±½		LSB
	AJ and LJ Suffix Parts			±½			±¾		
	ACJ and LCJ Suffix Parts				±½			±¾	
Differential Non-Linearity Max	Zero and Full-Scale Adjusted		±¼	±½		±½	±¾		
Monotonicity	AJ and LJ Suffix Parts			**12**			**12**		Bits
	ACJ and LCJ Suffix Parts			12	**12**		12	**12**	
Full-Scale (Gain) Error Max	R2 = 50Ω in *Figure 1*	5	±0.1	±0.20		±0.1	±0.20		% Full-Scale
Offset Error Max All Bits OFF, Logic "0"	Unipolar (*Figure 1* Pin 7 Open)	6	±0.01	±0.05		±0.01	±0.05		
	Bipolar (R1 and R2 = 50Ω in *Figure 2*)	7	±0.05	±0.1		±0.05	±0.15		
Zero Error Max MSB ON	Bipolar (R1 and R2 = 50Ω in *Figure 2*)	8	±0.05	±0.1		±0.05	±0.15		
Gain Adjustment Range Min	R2 = 50Ω ± 50Ω in *Figure 1*			±0.2			±0.2		
Bipolar Offset Adjustment Range Min	R1 = 50Ω ± 50Ω and R2 = 50Ω in *Figure 2*			±0.15			±0.15		
Full-Scale (Gain) Temperature Coefficients Max	AJ and LJ Suffix	9	1	**3**		5	**10**		ppm/°C
	ACJ and LCJ Suffix		1		**3**	5		**10**	
Unipolar Offset Temperature Coefficients Max	AJ and LJ Suffix		1	**2**		1	**2**		
	ACJ and LCJ Suffix		1		**2**	1		**2**	
Bipolar Zero Temperature Coefficients Max	AJ and LJ Suffix		5	**10**		5	**10**		
	ACJ and LCJ Suffix		5		**10**	5		**10**	
Output Resistance	Exclusive of Offset and Range R_S		7.5	6 to 10		7.5	6 to 10		kΩ
Current Output	Unipolar		−2	−1.6 to −2.4		−2	−1.6 to −2.4		mA
	Bipolar		±1.0	±0.8 to ±1.2		±1.0	±0.8 to ±1.2		

Electrical Characteristics (Continued) $V_{SUPPLY} = -15V \pm 5\%$ and $V_{REF} = 10.000V$ unless otherwise noted.

Boldface limits apply over temperature, $T_{MIN} \leq T_A \leq T_{MAX}$. For all other limits $T_A = 25°C$.

Parameter	Conditions	See Note	DAC1266A			DAC1266			Units
			Typ	Tested Limit (Note 2)	Design Limit (Note 3)	Typ	Tested Limit (Note 2)	Design Limit (Note 3)	
Output Capacitance			25			25			pF
Typ Output Voltage Ranges	Using Internal Offset and Range R_S		±2.5, ±5, ±10, 0 to 5, 0 to 10						V
Reference Input Resistance			20.8	15 to 25		20.8	15 to 25		kΩ
Output Compliance Voltage					**−1.5 to 10**			**−1.5 to 10**	V
DIGITAL AND DC CHARACTERISTICS									
Logic Input Voltage	Logic High AJ and LJ Suffix Bit ON ACJ and LCJ Suffix			**2 to 5.5** 1.9 to 5.5	**2 to 5.5**		**2 to 5.5** 1.9 to 5.5	**2 to 5.5**	V
Max	Logic Low AJ and LJ Suffix Bit OFF ACJ and LCJ Suffix			**0.8** 1.0	**0.8**		**0.8** 1.0	**0.8**	
Logic Input Current Max	Logic High AJ and LJ Suffix ACJ and LCJ Sufix		150 150	**300** 280	**300**	150 150	**300** 280	**300**	µA
	Logic Low AJ and LJ Suffix ACJ and LCJ Suffix		45 45	**100** 90	**100**	45 45	**100** 90	**100**	
Power Supply Current Max	V^- Supply = −15V ± 10%		−12	−18		−12	−18		mA
Power Dissipation Max	V^- Supply = −15V		180	270		180	270		mW
Power Supply Sensitivity Max	V^- Supply = −12V ± 5%	10	±15	±25		±15	±25		ppm of FS/ % V_{SUPPLY}
	V^- Supply = −15V ± 10%	10	±15	±25		±15	±25		
AC CHARACTERISTICS									
Settling Time Max	FSR Change		200		400	200		400	ns
Full-scale Transition Max	Delay Plus 10% to 90% Rise Time		15		30	15		30	ns
	Delay Plus 90% to 10% Fall Time		30		50	30		50	

Note 1: The typical θ_{JA} of the 24-pin package is 80° C/W.

Note 2: Tested and guaranteed to National's AOQL (Average Outgoing Quality Level).

Note 3: Guaranteed, but not 100% production tested. These limits are not used to calculate outgoing quality levels.

Note 4: Linearity error $= \dfrac{V_{OUT} - V_{OFFSET} - (D \times V_{LSB})}{V_{LSB}}$ where $V_{LSB} = \dfrac{V_{FS} - V_{OFFSET}}{4095}$ and D is the digital input (0 to 4095) which produced V_{OUT}.

Note 5: Percent gain error for 10V range $= \dfrac{(V_{FS} - V_{OFFSET}) - (4095/4096)V_{REF}}{V_{REF}} \times 100$.

Note 6: Unipolar offset error for 10V range $= (V_{OUT}/V_{REF}) \times 100$ in percent of full-scale.

Note 7: Bipolar offset error for 10V range $= \dfrac{V_{OUT} - (-V_{REF}/2)}{V_{REF}} \times 100$ in percent of full-scale.

Note 8: Bipolar zero error for 10V range $= (V_{OUT}/V_{REF}) \times 100$ in percent of full-scale.

Note 9: Gain error tempco $= \dfrac{(V_{FS} - V_{OFFSET}) \text{ at } (T_{MAX} \text{ or } T_{MIN}) - (V_{FS} - V_{OFFSET}) \text{ at } 25°C}{10V \text{ range} \times (T_{MAX} \text{ or } T_{MIN} - 25°C)} \times 10^6$ in ppm/°C.

Note 10: Power supply sensitivity for 10V range $= 10^6 \times \dfrac{(V_{FS} - V_{OFFSET}) \text{ at } (-13.5V) - (V_{FS} - V_{OFFSET}) \text{ at } (-16.5V)}{V_{REF} \times 20\%}$ in ppm of FS/% V_S.

Note 11: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 12: Human body model, 100 pF discharged through a 1.5 kΩ resistor.

Functional Description and Applications

1.0 BUFFERED VOLTAGE OUTPUT CONNECTION

The standard current-to-voltage conversion connections using an operational amplifier are shown here with the preferred trimming techniques. If a low offset operational amplifier (LF401A) is used, excellent performance can be obtained in many situations without trimming (an op amp with less than 0.5 mV maximum offset voltage should be used to keep offset errors below ½ LSB). Unipolar zero will typically be within ±½ LSB (plus op amp offset), and if a 50Ω fixed resistor is substituted for the 100Ω trimmer (R2, *Figure 1*), full-scale accuracy will be within 0.1% (0.20% maximum). Substituting a 50Ω resistor for the 100Ω bipolar offset trimmer (R1, *Figure 2*) will give a bipolar zero error typically within ±2 LSB (0.05%).

1.1 Unipolar Configuration (*Figure 1*)

This configuration will provide a unipolar 0V to 9.9976V output range.

Step 1—Offset Adjust (Zero)

Turn all bits OFF and adjust zero trimmer, R1, until the output reads 0.000V (1 LSB = 2.44 mV). In most cases this trim is not needed.

Step 2—Gain Adjust

Turn all bits ON and adjust 100Ω gain trimmer, R2, until the output is 9.9976V (full-scale adjusted to 1 LSB less than nominal full-scale of 10.000V). If a 10.2375V full-scale is desired (exactly 2.5 mV/bit), insert a 120Ω resistor in series with the gain resistor at pin 10 to the op amp output or use the LH0071 voltage reference.

1.2 Bipolar Configuration (*Figure 2*)

This configuration will provide a bipolar output voltage from −5.000V to 4.9976V, with positive full-scale occurring with all bits ON (all 1s).

Step 1—Offset Adjust

Turn OFF all bits. Adjust 100Ω offset trimmer, R1, to give −5.000V output.

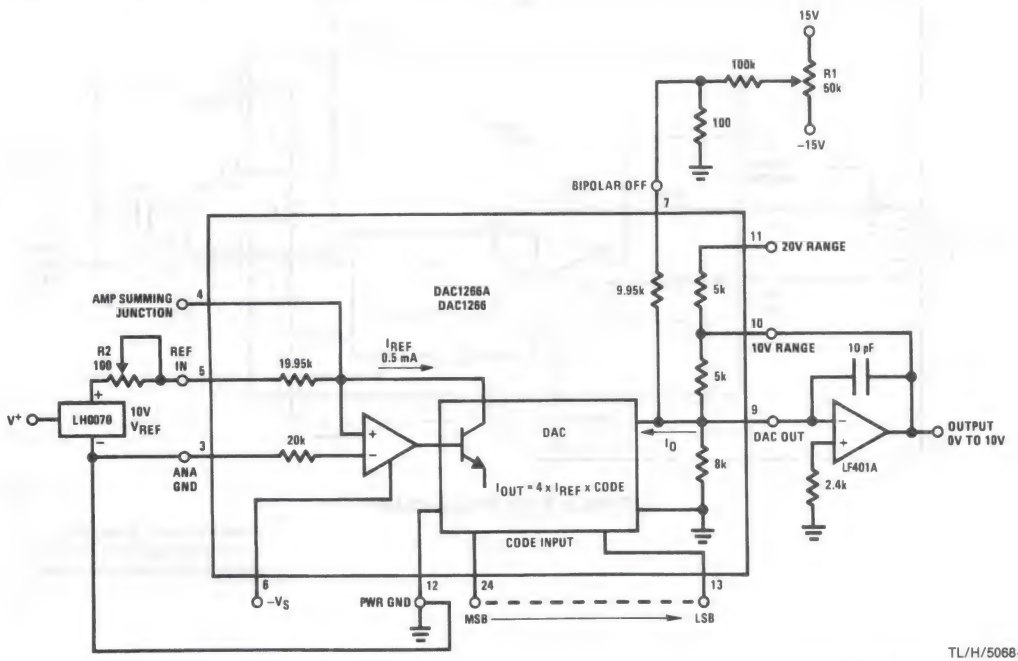
Step 2—Gain Adjust

Turn ON all bits. Adjust 100Ω gain trimmer, R2, to give a reading of 4.9976V.

Please note that it is not necessary to trim the op amp to obtain full accuracy at room temperature. In most bipolar situations, an op amp trim is unnecessary unless the untrimmed offset drift of the op amp is excessive. Bipolar zero error (MSB bit ON) is not adjusted separately and is typically < ±0.05% of FS after offset and gain adjust.

1.3 Other Voltage Ranges (*Figure 3*)

The DAC1266A and DAC1266 can also be easily configured for a unipolar 0V to 5V range or ±2.5V and ±10V bipolar ranges by using the additional 5k application resistor provided at the 20V range R terminal, pin 11. For a 5V span (0V to 5V or ±2.5V), the two 5k resistors are used in parallel by shorting pin 11 to pin 9 and connecting pin 10 to the op amp output and the bipolar offset either left open for unipolar or connected through a 100Ω pot to the external



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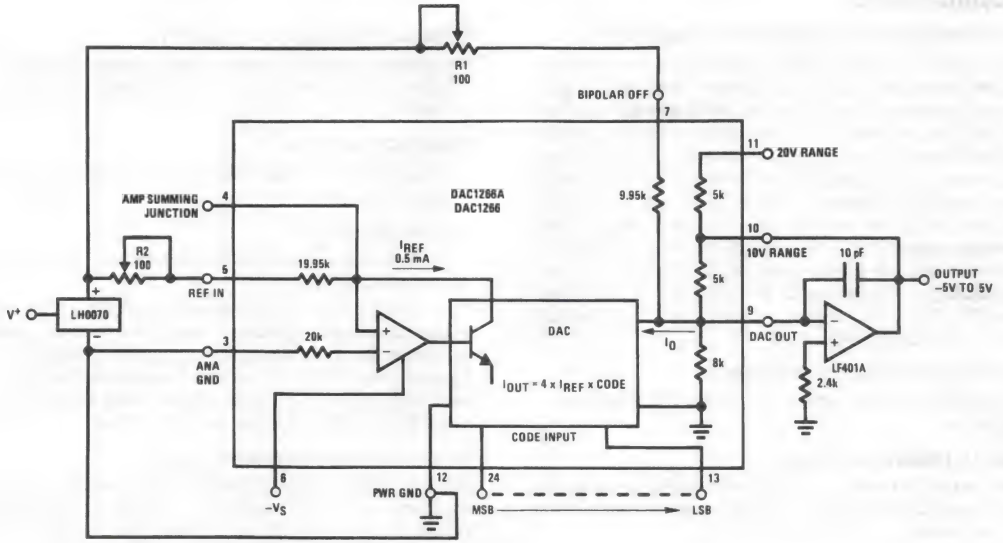
*Power and analog ground must have a common current return path. See section 3.0 for proper connections.

FIGURE 1. 0V to 10V Unipolar Voltage Output

Functional Description and Applications (Continued)



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*Power and analog ground must have a common current return path. See section 3.0 for proper connections.

FIGURE 2. ±5V Bipolar Voltage Output

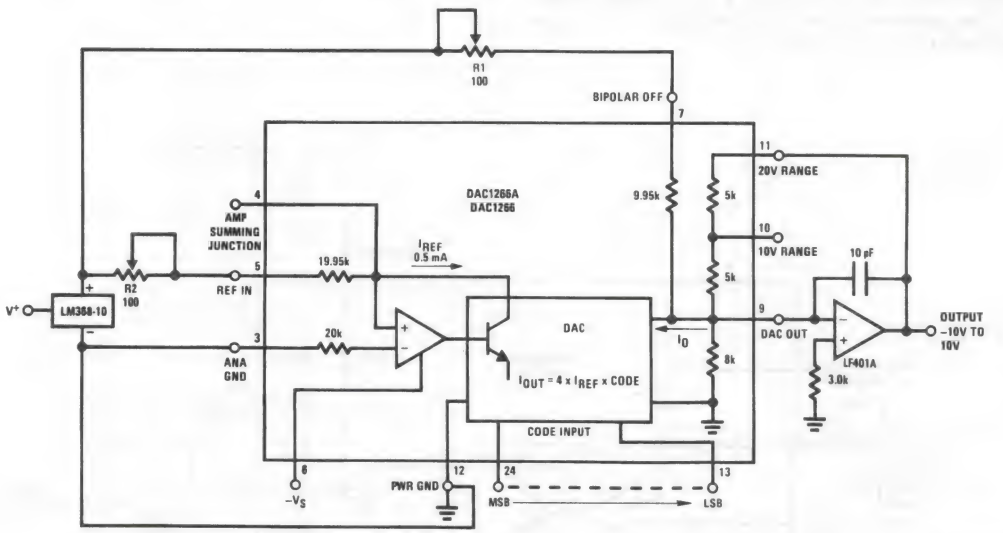

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FIGURE 3. ±10V Voltage Output

*Power and analog ground must have a common current return path. See section 3.0 for proper connections.

Functional Description and Applications (Continued)

reference for the bipolar range. For the ±10V range use the 5k resistors in series by connecting only pin 11 to the op amp output and connecting the bipolar offset as shown. The ±10V option is shown in *Figure 3*.

2.0 DIGITAL INPUT

The DAC1266A and DAC1266 use a standard positive true straight binary code for unipolar outputs (all 1s give full-scale output), and an offset binary code for bipolar output ranges. In the bipolar mode, with all 0s on the inputs, the output will go to negative full-scale; with 100...00 (only the MSB on), the output will be 0.00V; with all 1s, the output will go to positive full-scale.

The threshold of the digital input circuitry is set at 1.4V and does not vary with supply voltage. The input lines can interface with any type of 5V logic, TTL/DTL or CMOS, and have sufficiently low input currents to interface easily with unbuffered CMOS logic. The configuration of the input circuit is shown in *Figure 4*. The input line can be modelled as a 30 kΩ resistance connected to a −0.7V rail.

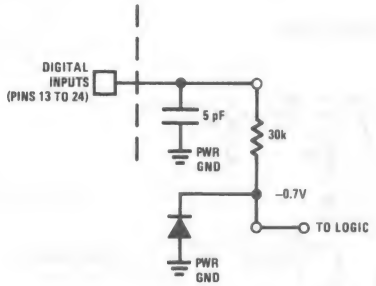

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FIGURE 4. Equivalent Digital Input Circuit

3.0 APPLICATION OF ANALOG AND POWER GROUND

The DAC1266A and DAC1266 have separate analog and power ground pins to allow optimum connections for low noise and high speed performance. The two ground lines can be separated by up to 200 mV without any loss in performance. There may be some loss in linearity beyond that level. If these DACs are to be used in a system in which the two grounds will be ultimately connected at some distance from the device, it is recommended that parallel back-to-back diodes be connected between the ground lines near the device to prevent a fault condition.

The analog ground at pin 3 is the ground reference point for the internal reference and is thus the "high quality" ground; it should be connected directly to the analog reference point of the system. The power ground at pin 12 can be connected to the most convenient ground reference point; analog power return is preferred, but digital ground is acceptable. If power ground contains high frequency noise beyond 200 mV, this noise may feed through the converter, so that some caution will be required in applying these grounds.

4.0 OUTPUT VOLTAGE COMPLIANCE

The DAC1266A and DAC1266 have a typical output compliance range from −2V to 10V. The current-steering output stages will be unaffected by changes in the output terminal voltage over that range. However, there is an equivalent output impedance of 8k in parallel with 25 pF at the output terminal which produces an equivalent error current if the voltage deviates from power ground. This is a linear effect that does not change with input code. Operation beyond the compliance limits may cause either output stage saturation or breakdown which results in non-linear performance. Compliance limits are a function of output current and negative supply.

5.0 DIRECT UNBUFFERED VOLTAGE OUTPUT FOR CABLE DRIVING

The wide compliance range allows direct current-to-voltage conversion with just an output resistor. *Figure 5* shows a connection using the gain and bipolar output resistors to give a ±1.60V bipolar swing. In this situation, the digital code is complementary binary. Other combinations of internal and external output resistors (R_X) can be used to scale to alternate voltage ranges, simply by appropriately scaling the 0 mA to −2 mA unipolar output current and using the 10.0V reference voltage for bipolar offset. For example, setting $R_X = 2.67$ kΩ gives a ±1V range with a 1 kΩ equivalent output impedance.

This connection is especially useful for directly driving a long cable at high speed. Using a 50Ω resistor for R_X would allow interface to a 50Ω cable with a ±50 mV full-scale swing.

6.0 HIGH SPEED 12-BIT A/D CONVERTERS

The fast settling characteristics of the DAC1266A and DAC1266 make them ideal for high speed successive approximation A/D converters. Shown in *Figure 6* is a configuration using standard components; this system completes a full 12-bit conversion in 10 μs unipolar or bipolar. This converter will be accurate to ±½ LSB of 12 bits and have a typical gain TC of 10 ppm/°C.

Functional Description and Applications (Continued)

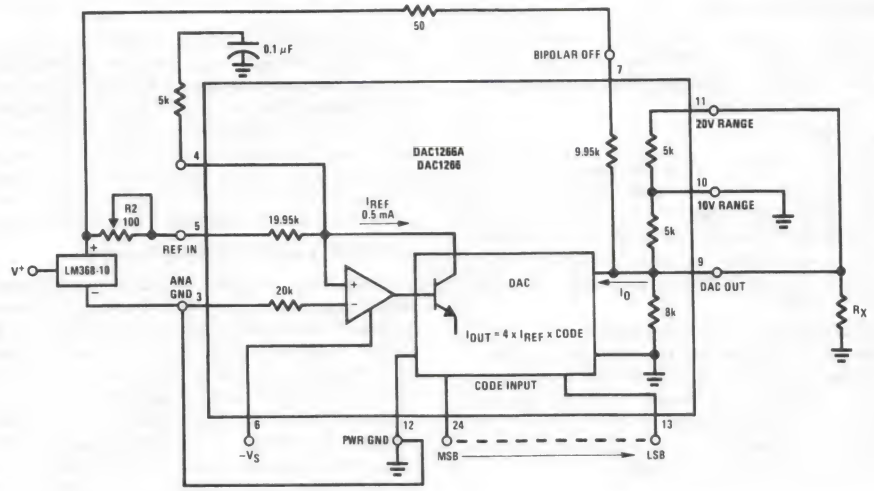
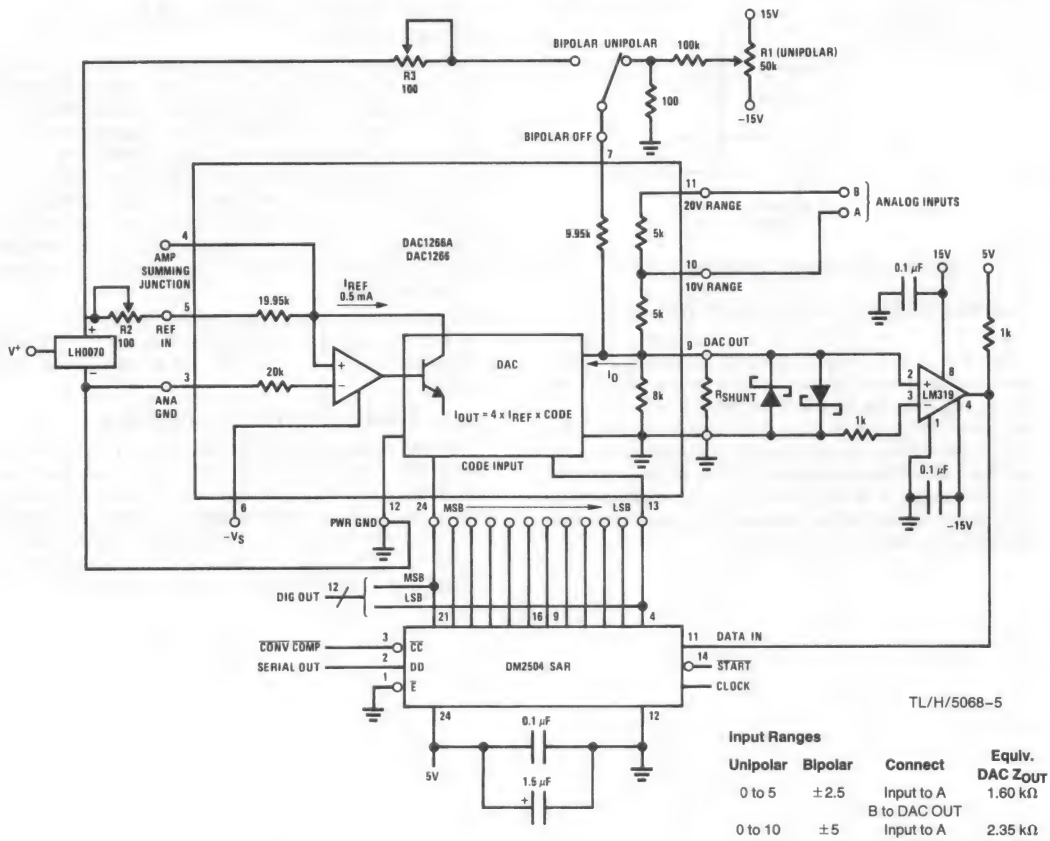


FIGURE 5. Unbuffered Bipolar Voltage Output



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Input Ranges

Unipolar	Bipolar	Connect	Equiv. DAC Z_{OUT}
0 to 5	±2.5	Input to A B to DAC OUT	1.60 kΩ
0 to 10	±5	Input to A	2.35 kΩ
0 to 20	±10	Input to B	3.08 kΩ

FIGURE 6. Fast Precision Analog to Digital Converter

Functional Description and Applications (Continued)

In the unipolar mode, the system range is 0V to 9.9976V, with each bit having a value of 2.44 mV. For true conversion accuracy, an A/D converter should be trimmed so that a given output code results from input levels from ½ LSB below to ½ LSB above the exact voltage represented by that code. Therefore, the converter zero point should be trimmed with an input voltage of 1.22 mV; trim R1 until the LSB just begins to appear in the output code (all other bits "0"). For full-scale, use an input voltage of 9.9963V (10V-1 LSB-½ LSB); then trim R2 until the LSB just begins to appear (all other bits "1").

The bipolar signal range is −5.0V to 4.9976V. Bipolar offset trimming is done by applying a −4.9988V input signal and trimming R3 for the LSB transition (all other bits "0").

Full-scale is set by applying a 4.9963V and trimming R2 for the LSB transition (all other bits "1"). In many applications, the pretrimmed internal resistors are sufficiently accurate that external trimmers will be unnecessary, especially in situations requiring less than full 12-bit ±½ LSB accuracy.

For fastest operation, the impedance at the comparator summing node must be minimized. However, lowering the impedance will reduce the voltage signal to the comparator (at an equivalent impedance at the summing node of 1 kΩ, 1 LSB = 0.5 mV), to the point that comparator performance will be sacrificed. The contribution to this impedance from the DAC will vary with the input configuration (*Figure 6*, Input Ranges Table).

To prevent dynamic errors, the input signal should have a low dynamic source impedance, such as that of the LF411A op amp.

Definition of Terms

Digital Inputs: The DAC1266A and DAC1266 accept digital input codes in binary format and may be user connected for any one of three binary codes: straight binary, two's complement, or offset binary.

Digital Input MSB LSB	Analog Output		
	Straight Binary	Offset Binary	Two's Complement*
000...000	zero	−FS (Full-Scale)	zero
011...111	½ FS-1 LSB	zero-1 LSB	+FS-1 LSB
100...000	½ FS	zero	−FS
111...111	+FS-1 LSB	+FS-1 LSB	zero-1 LSB

*Invert MSB with external inverter to obtain Two's Complement coding

Linearity Error: Linearity Error of a D/A converter is an important measure of its accuracy. It describes the deviation from an ideal straight line transfer curve drawn between zero (all bits OFF) and full-scale (all bits ON).

Differential Non-Linearity: For a D/A converter, it is the difference between the actual output voltage change and the ideal (1 LSB) voltage change for a one-bit change in code. A differential non-linearity of ±1 LSB or less guarantees monotonicity; i.e., the output always increases and never decreases for an increasing input. It is guaranteed by testing the major carry transitions; i.e., 100...000 to 011...111 etc.

Settling Time: Settling time is the time required for the output to settle to within the specified error band for any input code transition. It is usually specified for a full-scale or major carry transition.

Gain Tempco: The change in full-scale analog output over the specified temperature range expressed in parts per million of full-scale per °C (ppm of FS/°C). Gain error is measured with respect to 25°C at high (T_{MAX}) and low (T_{MIN}) temperatures. Gain tempco is calculated for both high (T_{MAX}−25°C) and low (25°C−T_{MIN}) ranges by dividing the gain error by the respective change in temperature. The specification is the larger of the two representing worst-case drift.

Offset Tempco: The change in analog output with all bits OFF over the specified temperature expressed in parts per million of full-scale per °C (ppm of FS/°C). Offset error is measured with respect to 25°C at high (T_{MAX}) and low (T_{MIN}) temperatures. Offset tempco is calculated for both high (T_{MAX}−25°C) and low (25°C−T_{MIN}) ranges by dividing the offset error by the respective change in temperature. The specification given is the larger of the two, representing worst-case drift.

Power Supply Sensitivity: Power supply sensitivity is a measure of the change in gain and offset of the D/A converter resulting from a change in −15V supply. It is specified under DC conditions and expressed as parts per million of full-scale per percent of change in power supply (ppm of FS/%).

Ordering Information

Temperature Range		0°C to 70°C	−55°C to +125°C
Linearity Error Over Temperature	±½ Bit	DAC1266ACJ	DAC1266AJ
	±¾ Bit	DAC1266LCJ	DAC1266LJ

Section 5

Sample and Hold

Section 5 Contents

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National Semiconductor

Sample and Hold Definition of Terms

Acquisition Time: The time required to acquire a new analog input voltage with an output step of 10V. Note that acquisition time is not just the time required for the output to settle, but also includes the time required for all internal nodes to settle so that the output assumes the proper value when switched to the hold mode.

Aperture Time: The delay required between "Hold" command and an input analog transition, so that the transition does not affect the held output.

Dynamic Sampling Error: The error introduced into the held output due to a changing analog input at the time the hold command is given. Error is expressed in mV with a given hold capacitor value and input slew rate. Note that this error term occurs even for long sample times.

Gain Error: The ratio of output voltage swing to input voltage swing in the sample mode expressed as a percent difference.

Hold Settling Time: The time required for the output to settle within 1 mV of final value after the "hold" logic command.

Hold Step: The voltage step at the output of the sample and hold when switching from sample mode to hold mode with a steady (DC) analog input voltage. Logic swing is 5V.

National Semiconductor

Sample and Hold Selection Guide

	LH0023	LH0043	LH0053	LH4860	Units
Accuracy Gain/Offset Error	0.01	0.1	0.2	0.01	% Max
Offset Voltage	20	40	7	5	mV Max
Droop Rate (25°C)					mV/sec
C_S = 1000 pF		10	6	500 (Note 2)	
C_S = 10000 pF	1	1			
Acquisition Time (25°C)					µs
C_S = 1000 pF	N/A	10	4 (Note 1)	0.15 (Note 2)	
C_S = 10000 pF	50	30			
Aperture Time (25°C)	150	20	10	6	ns
Temperature Range	−55 to +125	−55 to +125	−55 to +125	−55 to +125	°C
Comment	Low Drift	Medium Speed	High Speed	12-Bit High Speed	

Note 1: C_S = 100 pF

Note 2: C_S is internal

	LF198A	LF398A	LF198	LF398	LF298	Units
Accuracy Gain/Offset Error	0.01	0.01	0.02	0.02	0.02	% Max
Offset Voltage	2	3	5	10	5	mV Max
Droop Rate (25°C)						mV/sec
C_S = 1000 pF	30	30	30	30	30	
C_S = 10000 pF	3	3	3	3	3	
Acquisition Time (25°C)						µs
C_S = 1000 pF	4	4	4	4	4	
C_S = 10000 pF	20	20	20	20	20	
Aperture Time (25°C)	250	250	250	250	250	ns
Temperature Range	−55 to +125	0 to +70	−55 to +125	0 to +70	−25 to +85	°C
Comment	Low Drift	Low Drift	General Purpose	General Purpose	Low Drift	

National Semiconductor

LF198/LF298/LF398, LF198A/LF398A Monolithic Sample and Hold Circuits

General Description

The LF198/LF298/LF398 are monolithic sample and hold circuits which utilize BI-FET technology to obtain ultra-high dc accuracy with fast acquisition of signal and low droop rate. Operating as a unity gain follower, dc gain accuracy is 0.002% typical and acquisition time is as low as 6 μs to 0.01%. A bipolar input stage is used to achieve low offset voltage and wide bandwidth. Input offset adjust is accomplished with a single pin, and does not degrade input offset drift. The wide bandwidth allows the LF198 to be included inside the feedback loop of 1 MHz op amps without having stability problems. Input impedance of $10^{10}\Omega$ allows high source impedances to be used without degrading accuracy.

P-channel junction FET's are combined with bipolar devices in the output amplifier to give droop rates as low as 5 mV/min with a 1 μF hold capacitor. The JFET's have much lower noise than MOS devices used in previous designs and do not exhibit high temperature instabilities. The overall design guarantees no feed-through from input to output in the hold mode, even for input signals equal to the supply voltages.

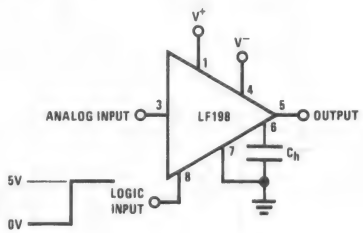
Features

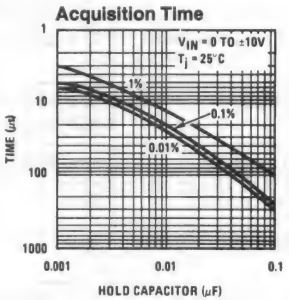
- Operates from $\pm$5V to $\pm$18V supplies
- Less than 10 μs acquisition time
- TTL, PMOS, CMOS compatible logic input
- 0.5 mV typical hold step at C_h = 0.01 μF
- Low input offset
- 0.002% gain accuracy
- Low output noise in hold mode
- Input characteristics do not change during hold mode
- High supply rejection ratio in sample or hold
- Wide bandwidth

Logic inputs on the LF198 are fully differential with low input current, allowing direct connection to TTL, PMOS, and CMOS. Differential threshold is 1.4V. The LF198 will operate from $\pm$5V to $\pm$18V supplies. It is available in an 8-lead TO-5 package.

An "A" version is available with tightened electrical specifications.

Typical Connection and Performance Curve





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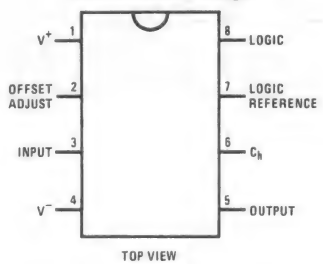
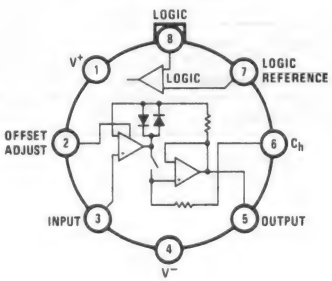
Connection Diagrams

Dual-In-Line Package



Order Number LF398N or LF398AN
See NS Package Number N08E

Metal Can Package



TL/H/5692-11

Order Number LF198H, LF298H, LF398H, LF198AH or LF398AH
See NS Package Number H08C

5

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	±18V
Power Dissipation (Package Limitation) (Note 1)	500 mW
Operating Ambient Temperature Range	
LF198/LF198A	−55°C to +125°C
LF298	−25°C to +85°C
LF398/LF398A	0°C to +70°C
Storage Temperature Range	−65°C to +150°C
Input Voltage	Equal to Supply Voltage
Logic To Logic Reference Differential Voltage (Note 2)	+7V, −30V
Output Short Circuit Duration	Indefinite
Hold Capacitor Short Circuit Duration	10 sec
Lead Temperature (Soldering, 10 seconds)	260°C
Thermal Resistance (θ_{JA}) (typicals)	
H package	215°C/W (Board mount in still air)
	85°C/W (Board mount in 400LF/min air flow)
N package	115°C/W
θ_{JC} (typical)	20°C/W

Electrical Characteristics (Note 3)

Parameter	Conditions	LF198/LF298			LF398			Units
		Min	Typ	Max	Min	Typ	Max	
Input Offset Voltage, (Note 6)	$T_j = 25°C$ Full Temperature Range		1	3 5		2	7 10	mV mV
Input Bias Current, (Note 6)	$T_j = 25°C$ Full Temperature Range		5	25 75		10	50 100	nA nA
Input Impedance	$T_j = 25°C$		10^{10}			10^{10}		Ω
Gain Error	$T_j = 25°C$, $R_L = 10k$ Full Temperature Range		0.002	0.005 0.02		0.004	0.01 0.02	% %
Feedthrough Attenuation Ratio at 1 kHz	$T_j = 25°C$, $C_h = 0.01\ \mu F$	86	96		80	90		dB
Output Impedance	$T_j = 25°C$, "HOLD" mode Full Temperature Range		0.5	2 4		0.5	4 6	Ω Ω
"HOLD" Step, (Note 4)	$T_j = 25°C$, $C_h = 0.01\ \mu F$, $V_{OUT} = 0$		0.5	2.0		1.0	2.5	mV
Supply Current, (Note 6)	$T_j \geq 25°C$		4.5	5.5		4.5	6.5	mA
Logic and Logic Reference Input Current	$T_j = 25°C$		2	10		2	10	μA
Leakage Current into Hold Capacitor (Note 6)	$T_j = 25°C$, (Note 5) Hold Mode		30	100		30	200	pA
Acquisition Time to 0.1%	$\Delta V_{OUT} = 10V$, $C_h = 1000$ pF $C_h = 0.01\ \mu F$		4 20			4 20		μs μs
Hold Capacitor Charging Current	$V_{IN} - V_{OUT} = 2V$		5			5		mA
Supply Voltage Rejection Ratio	$V_{OUT} = 0$	80	110		80	110		dB
Differential Logic Threshold	$T_j = 25°C$	0.8	1.4	2.4	0.8	1.4	2.4	V

Electrical Characteristics (Continued) (Note 3)

Parameter	Conditions	LF198A			LF398A			Units
		Min	Typ	Max	Min	Typ	Max	
Input Offset Voltage, (Note 6)	$T_j = 25°C$ Full Temperature Range		1	1 2		2	2 3	mV mV
Input Bias Current, (Note 6)	$T_j = 25°C$ Full Temperature Range		5	25 75		10	25 50	nA nA
Input Impedance	$T_j = 25°C$		10^{10}			10^{10}		Ω
Gain Error	$T_j = 25°C$, $R_L = 10k$ Full Temperature Range		0.002	0.005 0.01		0.004	0.005 0.01	% %
Feedthrough Attenuation Ratio at 1 kHz	$T_j = 25°C$, $C_h = 0.01\ \mu F$	86	96		86	90		dB
Output Impedance	$T_j = 25°C$, "HOLD" mode Full Temperature Range		0.5	1 4		0.5	1 6	Ω Ω
"HOLD" Step, (Note 4)	$T_j = 25°C$, $C_h = 0.01\mu F$, $V_{OUT} = 0$		0.5	1		1.0	1	mV
Supply Current, (Note 6)	$T_j \geq 25°C$		4.5	5.5		4.5	6.5	mA
Logic and Logic Reference Input Current	$T_j = 25°C$		2	10		2	10	μA
Leakage Current into Hold Capacitor (Note 6)	$T_j = 25°C$, (Note 5) Hold Mode		30	100		30	100	pA
Acquisition Time to 0.1%	$\Delta V_{OUT} = 10V$, $C_h = 1000$ pF $C_h = 0.01\ \mu F$		4 20	6 25		4 20	6 25	μs μs
Hold Capacitor Charging Current	$V_{IN} - V_{OUT} = 2V$		5			5		mA
Supply Voltage Rejection Ratio	$V_{OUT} = 0$	90	110		90	110		dB
Differential Logic Threshold	$T_j = 25°C$	0.8	1.4	2.4	0.8	1.4	2.4	V

Note 1: The maximum junction temperature of the LF198/LF198A is 150°C, for the LF298, 115°C, and for the LF398/LF398A, 100°C. When operating at elevated ambient temperature, the power dissipation must be derated based on a thermal resistance (Θ_{jA}) of 150°C/W.

Note 2: Although the differential voltage may not exceed the limits given, the common-mode voltage on the logic pins may be equal to the supply voltages without causing damage to the circuit. For proper logic operation, however, one of the logic pins must always be at least 2V below the positive supply and 3V above the negative supply.

Note 3: Unless otherwise specified, the following conditions apply. Unit is in "sample" mode, $V_S = \pm 15V$, $T_j = 25°C$, $-11.5V \leq V_{IN} \leq +11.5V$, $C_h = 0.01\ \mu F$, and $R_L = 10\ k\Omega$. Logic reference voltage = 0V and logic voltage = 2.5V.

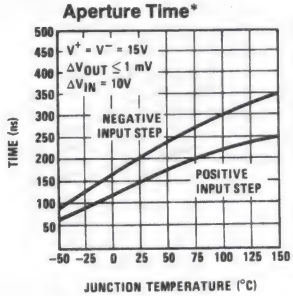
Note 4: Hold step is sensitive to stray capacitive coupling between input logic signals and the hold capacitor. 1 pF, for instance, will create an additional 0.5 mV step with a 5V logic swing and a 0.01μF hold capacitor. Magnitude of the hold step is inversely proportional to hold capacitor value.

Note 5: Leakage current is measured at a *junction* temperature of 25°C. The effects of junction temperature rise due to power dissipation or elevated ambient can be calculated by doubling the 25°C value for each 11°C increase in chip temperature. Leakage is guaranteed over full input signal range.

Note 6: These parameters guaranteed over a supply voltage range of ±5 to ±18V, and an input range of $-V_S + 3.5V \leq V_{IN} \leq +V_S - 3.5V$.

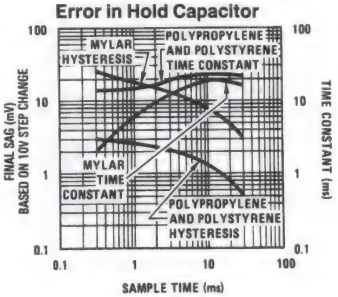
Typical Performance Characteristics

Aperture Time*

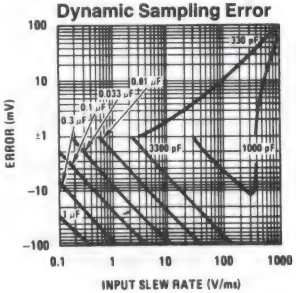


*See Definition of Terms

Dielectric Absorption Error in Hold Capacitor


MYLAR HYSTERESIS
POLYPROPYLENE AND POLYSTYRENE TIME CONSTANT
MYLAR TIME CONSTANT
POLYPROPYLENE AND POLYSTYRENE HYSTERESIS
FINAL SAG (mV) BASED ON 10V STEP CHANGE
TIME CONSTANT (ms)
SAMPLE TIME (ms)

Dynamic Sampling Error



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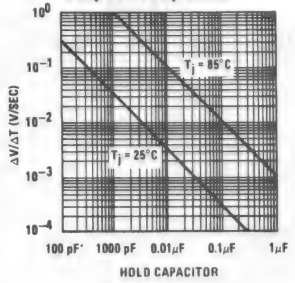
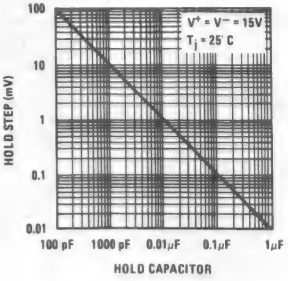
Typical Performance Characteristics (Continued)

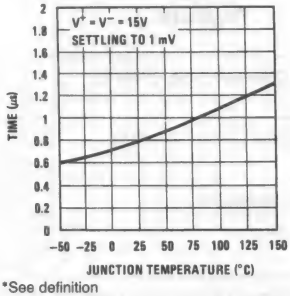
Output Droop Rate



Hold Step

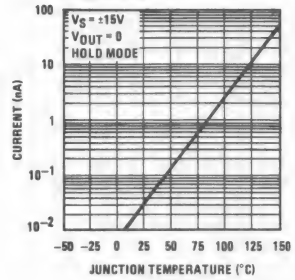


"Hold" Settling Time*

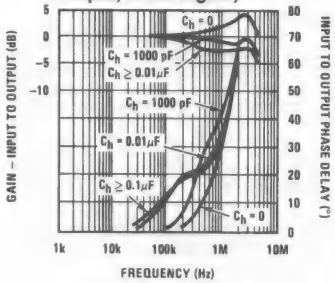


*See definition

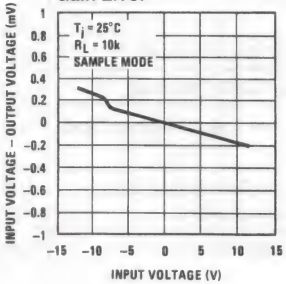
Leakage Current into Hold Capacitor



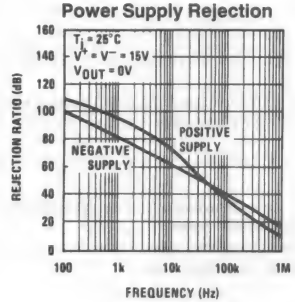
Phase and Gain (Input to Output, Small Signal)



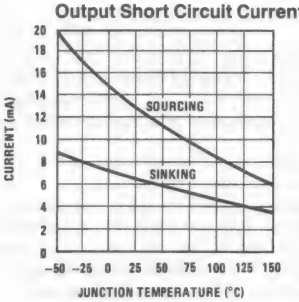
Gain Error



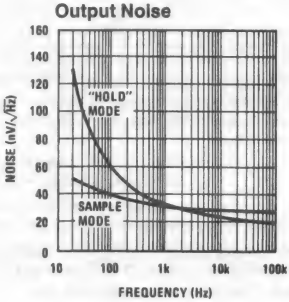
Power Supply Rejection



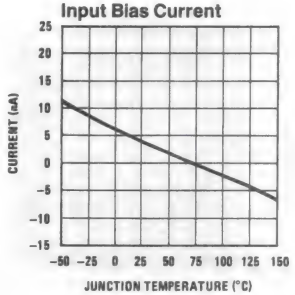
Output Short Circuit Current



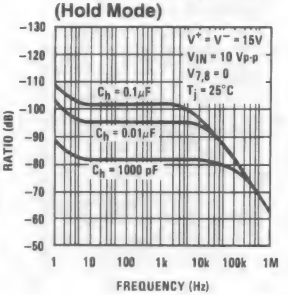
Output Noise



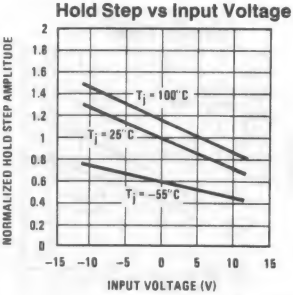
Input Bias Current



Feedthrough Rejection Ratio (Hold Mode)



Hold Step vs Input Voltage



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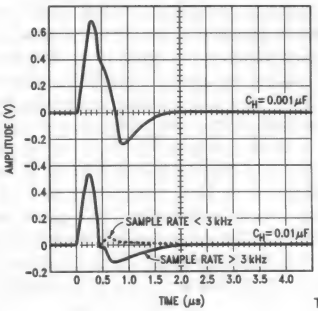
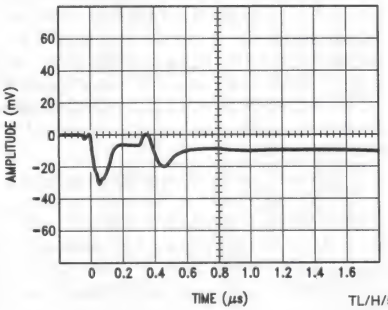
Typical Performance Characteristics (Continued)

Output Transient at Start of Sample Mode



TL/H/5692–12

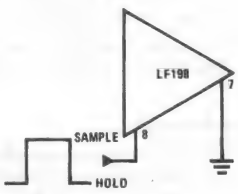
Output Transient at Start of Hold Mode



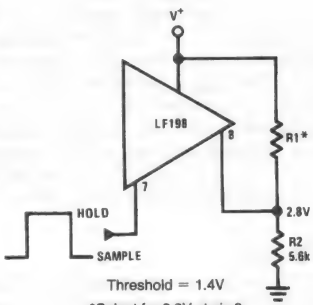
TL/H/5692–13

Logic Input Configurations

TTL & CMOS
3V ≤ V_L (Hi State) ≤ 7V

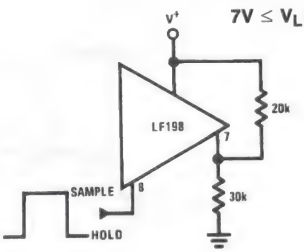


Threshold = 1.4V

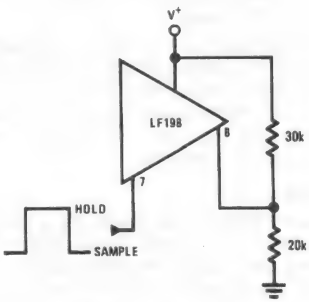


Threshold = 1.4V
*Select for 2.8V at pin 8

CMOS
7V ≤ V_L (Hi State) ≤ 15V

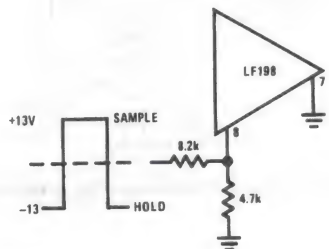


Threshold = 0.6 (V^+) + 1.4V

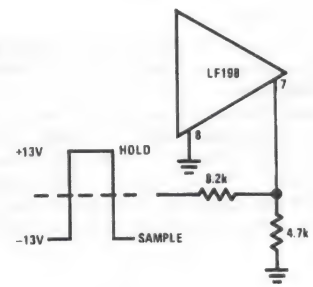


Threshold = 0.6 (V^+) − 1.4V

Op Amp Drive



Threshold ≈ +4V



Threshold = −4V

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5

Application Hints

Hold Capacitor

Hold step, acquisition time, and droop rate are the major trade-offs in the selection of a hold capacitor value. Size and cost may also become important for larger values. Use of the curves included with this data sheet should be helpful in selecting a reasonable value of capacitance. Keep in mind that for fast repetition rates or tracking fast signals, the capacitor drive currents may cause a significant temperature rise in the LF198.

A significant source of error in an accurate sample and hold circuit is dielectric absorption in the hold capacitor. A mylar cap, for instance, may "sag back" up to 0.2% after a quick change in voltage. A long sample time is required before the circuit can be put back into the hold mode with this type of capacitor. Dielectrics with very low hysteresis are polystyrene, polypropylene, and Teflon. Other types such as mica and polycarbonate are not nearly as good. Ceramic is unusable with $> 1\%$ hysteresis. The advantage of polypropylene over polystyrene is that it extends the maximum ambient temperature from 85°C to 100°C. "NPO" or "COG" capacitors are now available for 125°C operation and also have low dielectric absorption. For more exact data, see the curve *Dielectric Absorption Error.* The hysteresis numbers on the curve are final values, taken after full relaxation. The hysteresis error can be significantly reduced if the output of the LF198 is digitized quickly after the hold mode is initiated. The hysteresis relaxation time constant in polypropylene, for instance, is 10—50 ms. If A-to-D conversion can be made within 1 ms, hysteresis error will be reduced by a factor of ten.

DC and AC Zeroing

DC zeroing is accomplished by connecting the offset adjust pin to the wiper of a 1 kΩ potentiometer which has one end tied to V^+ and the other end tied through a resistor to ground. The resistor should be selected to give ≈ 0.6 mA through the 1k potentiometer.

AC zeroing (hold step zeroing) can be obtained by adding an inverter with the adjustment pot tied input to output. A 10 pF capacitor from the wiper to the hold capacitor will give ±4 mV hold step adjustment with a 0.01 μF hold capacitor and 5V logic supply. For larger logic swings, a smaller capacitor (< 10 pF) may be used.

Logic Rise Time

For proper operation, logic signals into the LF198 must have a minimum dV/dt of 1.0 V/μs. Slower signals will cause excessive hold step. If a R/C network is used in front of the logic input for signal delay, calculate the slope of the waveform at the threshold point to ensure that it is at least 1.0 V/μs.

Sampling Dynamic Signals

Sample error to moving input signals probably causes more confusion among sample-and-hold users than any other parameter. The primary reason for this is that many users make the assumption that the sample and hold amplifier is truly locked on to the input signal while in the sample mode. In actuality, there are finite phase delays through the circuit creating an input-output differential for fast moving signals. In addition, although the output may have settled, the hold capacitor has an additional lag due to the 300Ω series resistor on the chip. This means that at the moment the "hold" command arrives, the hold capacitor voltage may be somewhat different than the actual analog input. The effect of these delays is opposite to the effect created by delays in the logic which switches the circuit from sample to hold. For example, consider an analog input of 20 Vp-p at 10 kHz. Maximum dV/dt is 0.6 V/μs. With no analog phase delay and 100 ns logic delay, one could expect up to (0.1 μs) (0.6V/μs) = 60 mV error if the "hold" signal arrived near maximum dV/dt of the input. A positive-going input would give a +60 mV error. Now assume a 1 MHz (3 dB) bandwidth for the overall analog loop. This generates a phase delay of 160 ns. If the hold capacitor sees this exact delay, then error due to analog delay will be (0.16 μs) (0.6 V/μs) = −96 mV. Total output error is +60 mV (digital) −96 mV (analog) for a total of −36 mV. To add to the confusion, analog delay is proportioned to hold capacitor value while digital delay remains constant. A family of curves (dynamic sampling error) is included to help estimate errors.

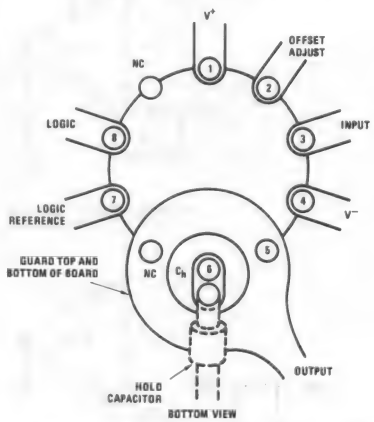
A curve labeled *Aperture Time* has been included for sampling conditions where the input is steady during the sampling period, but may experience a sudden change nearly coincident with the "hold" command. This curve is based on a 1 mV error fed into the output.

A second curve, *Hold Settling Time* indicates the time required for the output to settle to 1 mV after the "hold" command.

Digital Feedthrough

Fast rise time logic signals can cause hold errors by feeding externally into the analog input at the same time the amplifier is put into the hold mode. To minimize this problem, board layout should keep logic lines as far as possible from the analog input. Grounded guarding traces may also be used around the input line, especially if it is driven from a high impedance source. Reducing high amplitude logic signals to 2.5V will also help.

Guarding Technique



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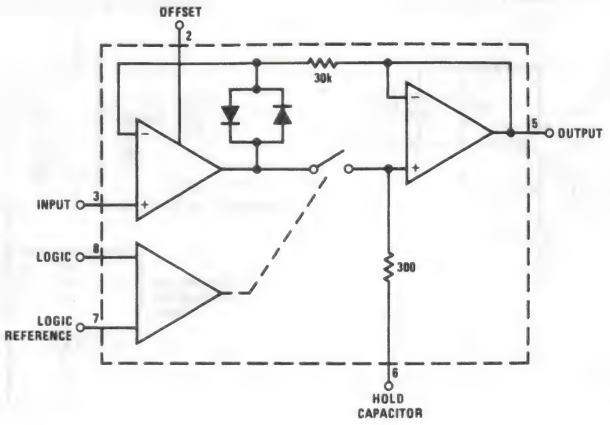
Use 10-pin layout. Guard around C_h is tied to output.

Functional Diagram



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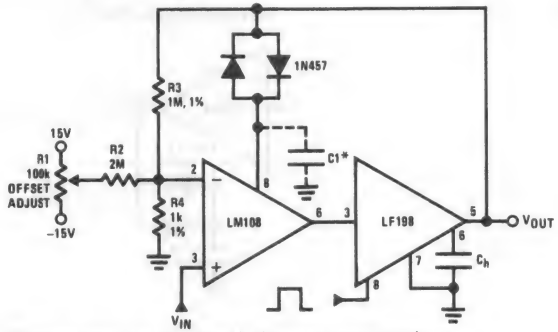
Typical Applications (Continued)

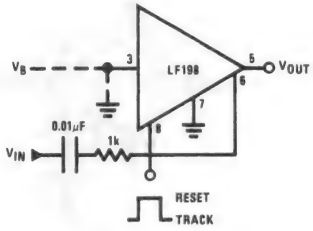
X1000 Sample & Hold



*For lower gains, the LM108 must be frequency compensated

Use $\approx \frac{100}{A_V}$ pF from comp 2 to ground

Sample and Difference Circuit
(Output Follows Input in *Hold* Mode)



$V_{OUT} = V_B + \Delta V_{IN}$(HOLD MODE)

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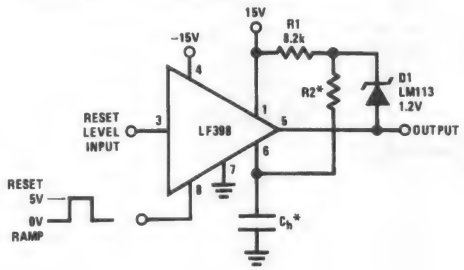
Typical Applications (Continued)

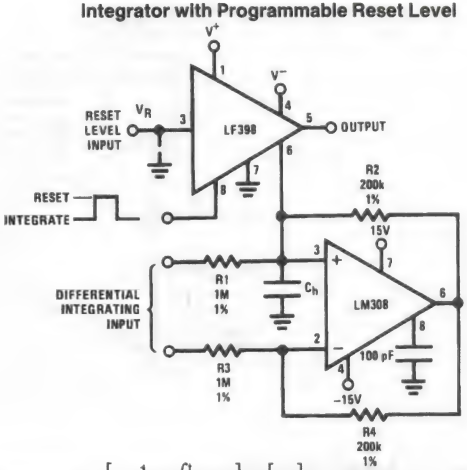
Ramp Generator with Variable Reset Level



*Select for ramp rate R2 ≥ 10k

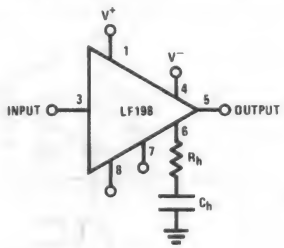
$$\frac{\Delta V}{\Delta T} = \frac{1.2V}{(R2)\,(C_h)}$$

Integrator with Programmable Reset Level



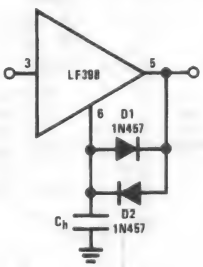
$$V_{OUT}\ (\text{Hold Mode}) = \left[\frac{1}{(R1)\,(C_h)}\int_0^t V_{IN}\,dt\right] + \left[V_R\right]$$

Output Holds at Average of Sampled Input

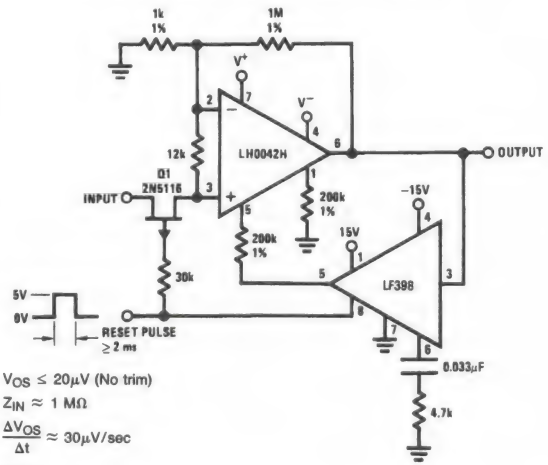


$$\text{Select } (R_h)\,(C_h) \geq \frac{1}{2\pi f_{IN}\ (\text{Min})}$$

Increased Slew Current



Reset Stabilized Amplifier (Gain of 1000)



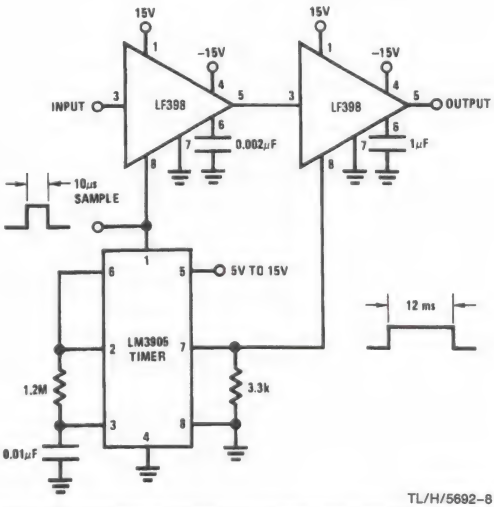
$V_{OS} \leq 20\mu V$ (No trim)

$Z_{IN} \approx 1\ M\Omega$

$\frac{\Delta V_{OS}}{\Delta t} \approx 30\mu V/\text{sec}$

$\frac{\Delta V_{OS}}{\Delta T} \approx 0.1\mu V/°C$

Fast Acquisition, Low Droop Sample & Hold



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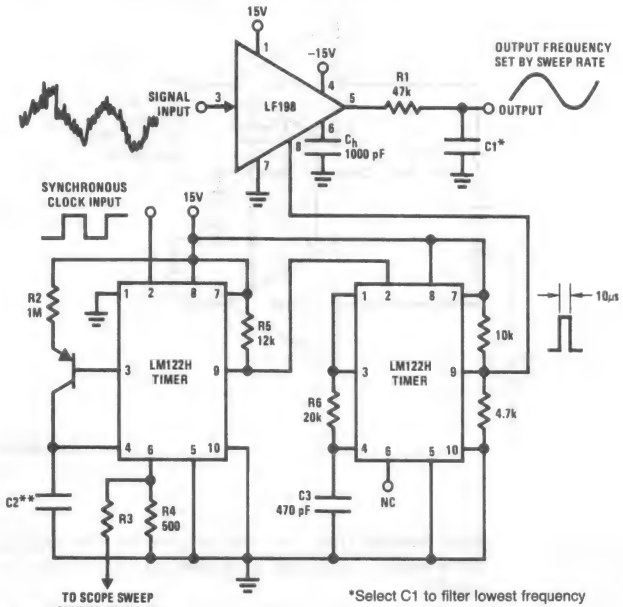
Typical Applications (Continued)

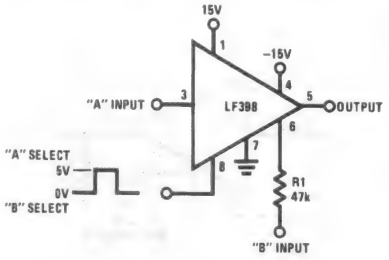
Synchronous Correlator for Recovering Signals Below Noise Level



*Select C1 to filter lowest frequency component of input noise

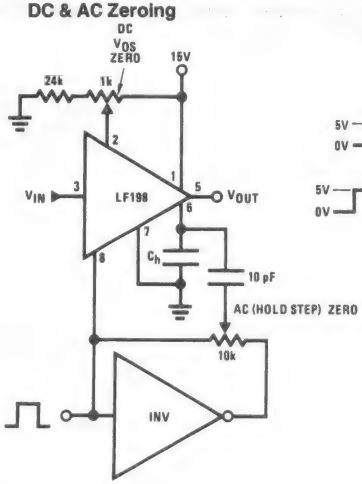
**Select C2 @ $\approx 5 \times 10^{-6}/f_{IN}$

2-Channel Switch

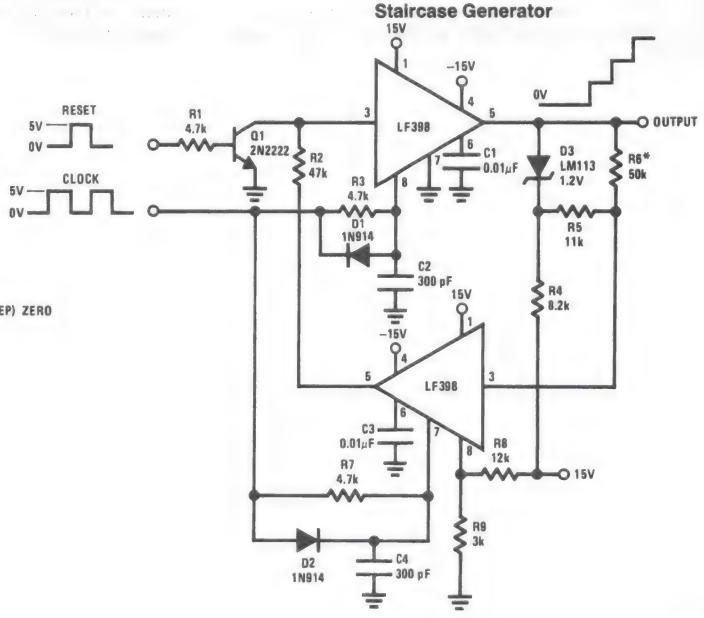


	A	B
Gain	1 ± 0.02%	1 ± 0.2%
Z_{IN}	$10^{10}\Omega$	47 kΩ
BW	≅ 1 MHz	≅ 400 kHz
Crosstalk @ 1 kHz	−90 dB	−90 dB
Offset	≤ 6 mV	≤ 75 mV

DC & AC Zeroing



Staircase Generator



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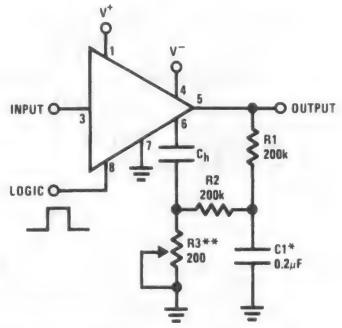
*Select for step height
50k ⟶ ≅ 1V Step

Typical Applications (Continued)

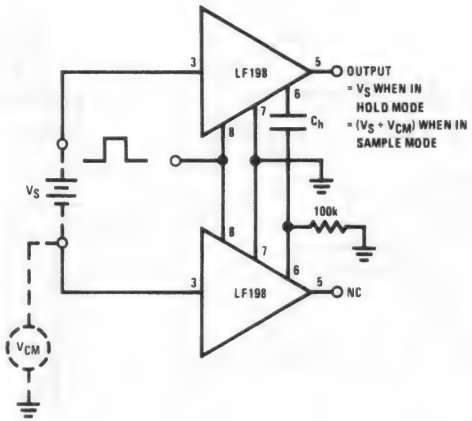
Capacitor Hysteresis Compensation



*Select for time constant C1 = $\frac{\tau}{100k}$

**Adjust for amplitude

Differential Hold



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Definition of Terms

Hold Step: The voltage step at the output of the sample and hold when switching from sample mode to hold mode with a steady (dc) analog input voltage. Logic swing is 5V.

Acquisition Time: The time required to acquire a new analog input voltage with an output step of 10V. Note that acquisition time is not just the time required for the output to settle, but also includes the time required for all internal nodes to settle so that the output assumes the proper value when switched to the hold mode.

Gain Error: The ratio of output voltage swing to input voltage swing in the sample mode expressed as a per cent difference.

Hold Settling Time: The time required for the output to settle within 1 mV of final value after the "hold" logic command.

Dynamic Sampling Error: The error introduced into the held output due to a changing analog input at the time the hold command is given. Error is expressed in mV with a given hold capacitor value and input slew rate. Note that this error term occurs even for long sample times.

Aperture Time: The delay required between "Hold" command and an input analog transition, so that the transition does not affect the held output.

National Semiconductor

PRELIMINARY

LF13006/LF13007 Digital Gain Set

General Description

The LF13006 and LF13007 are precision digital gain sets used for accurately setting non-inverting op amp gains. Gains are set with a 3-bit digital word which can be latched in with $\overline{WR}$ and $\overline{CS}$ pins. All digital inputs are TTL and CMOS compatible.

The LF13006 shown below will set binary scaled gains of 1, 2, 4, 8, 16, 32, 64, and 128. The LF13007 will set gains of 1, 2, 5, 10, 20, 50, and 100 (a common attenuator sequence). In addition, both versions have several taps and two uncommitted matching resistors that allow customization of the gain.

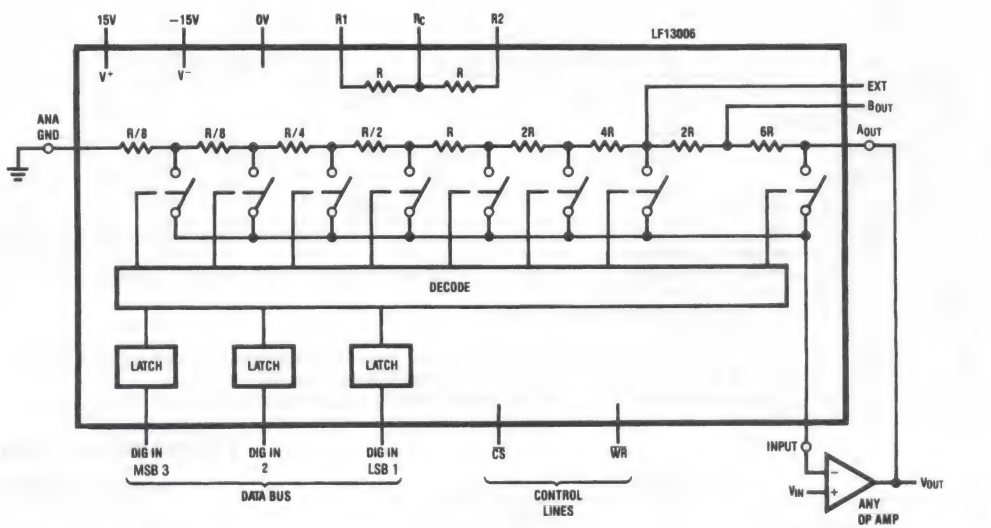
The gains are set with precision thin film resistors. The low temperature coefficient of the thin film resistors and their excellent tracking result in gain ratios which are virtually independent of temperature.

The LF13006, LF13007 used in conjunction with an amplifier not only satisfies the need for a digitally programmable amplifier in microprocessor based systems, but is also useful for discrete applications, eliminating the need to find 0.5% resistors in the ratio of 100 to 1 which track each other over temperature.

Features

- TTL and CMOS compatible logic levels
- Microprocessor compatible
- Gain error 0.5% max
- Binary or scope knob gains
- Wide supply range + 5V to ± 18V
- Packaged in 16-pin DIP

Block Diagram and Typical Application (LF13006)



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Note: R ≅ 15 kΩ

Order Number LF13006N or LF13007N
See NS Package Number N16A

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage, V^+ to V^-	36V
Supply Voltage, V^+ to GND	25V
Voltage at Any Digital Input	V^+ to GND
Analog Voltage	V^+ to (V^- + 2V)

Operating Ratings (Note 1)

Operating Temperature Range	−40°C to +85°C
Lead Temp. (Soldering, 10 seconds)	260°C

Electrical Characteristics (Note 2)

Parameter	Conditions	Typ (Note 3)	Tested Limit (Note 4)	Design Limit (Note 5)	Units
Gain Error	A_{OUT} = ±10V ANA GND = 0V I_{INPUT} < 10 nA	0.3	0.5	**0.5**	%(max)
Gain Temperature Coefficient	A_{OUT} = ±10V ANA GND = 0V	0.001			%/°C
Digital Input Voltage Low High		 1.4 1.6	 0.8 2.0	 **0.8** **2.0**	 V(max) V(min)
Digital Input Current Low High	 V_{IL} = 0V V_{IH} = 5V	 −38 0.0001	 −100 1	 **−100** **1**	 μA(max) μA(max)
Positive Power Supply Current	All Logic Inputs Low	2	5	**5**	mA(max)
Negative Power Supply Current	All Logic Inputs Low	−1.7	−5	**−5**	mA(max)
$\overline{\text{Write}}$ Pulse Width, t_W	V_{IL} = 0V, V_{IH} = 5V		150		ns(min)
$\overline{\text{Chip Select}}$ Set-Up Time, t_{CS}	V_{IL} = 0V, V_{IH} = 5V		250		ns(min)
$\overline{\text{Chip Select}}$ Hold Time, t_{CH}	V_{IL} = 0V, V_{IH} = 5V		0		ns(min)
DIG IN Set-Up Time, t_{DS}	V_{IL} = 0V, V_{IH} = 5V		150		ns(min)
DIG IN Hold Time, t_{DH}	V_{IL} = 0V, V_{IH} = 5V		60		ns(min)
Switching Time for Gain Change	(Note 4)	200			ns(max)
Switch On Resistance		3			kΩ
Unit Resistance, R		15	12–18		kΩ
R1 and R2 Mismatch		0.3	0.5	**0.5**	%(max)
R1/R2 Temperature Coefficient		0.001			%/°C

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its specified operating conditions.

Note 2: Parameters are specified at V^+ = 15V and V^- = −15V. Min V^+ to ground voltage is 5V. Min V^+ to V^- voltage is 5V. **Boldface numbers apply over full operating temperature ranges.** All other numbers apply at $T_A = T_j = 25°C$.

Note 3: Typicals are at 25°C and represent most likely parametric norm.

Note 4: Guaranteed and 100% production tested.

Note 5: Guaranteed (but not 100% production tested) over the operating temperature. These limits are not used to calculate outgoing quality levels.

Note 6: Settling time for gain change is the switching time for gain change plus settling time (see section on Settling Time).

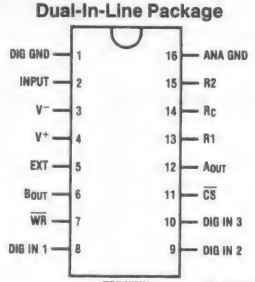
Note 7: $\overline{WR}$ minimum high threshold voltage increases to 2.4V under the extreme conditions when all three digital inputs are simultaneously taken from 0V to 5V at a slew rate of greater than 500V/μS.

GAIN TABLE

Digital Input			Gain			
			LF13006		LF13007	
DIG In 3	DIG In 2	DIG In 1	A_{OUT}	B_{OUT}	A_{OUT}	B_{OUT}
0	0	0	1	1	1	1
0	0	1	2	1.25	1.25	1
0	1	0	4	2.5	2	1.6
0	1	1	8	5	5	4
1	0	0	16	10	10	8
1	0	1	32	20	20	16
1	1	0	64	40	50	40
1	1	1	128	80	100	80

Connection Diagram

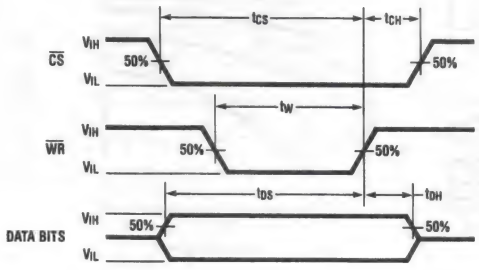
Dual-In-Line Package



TOP VIEW

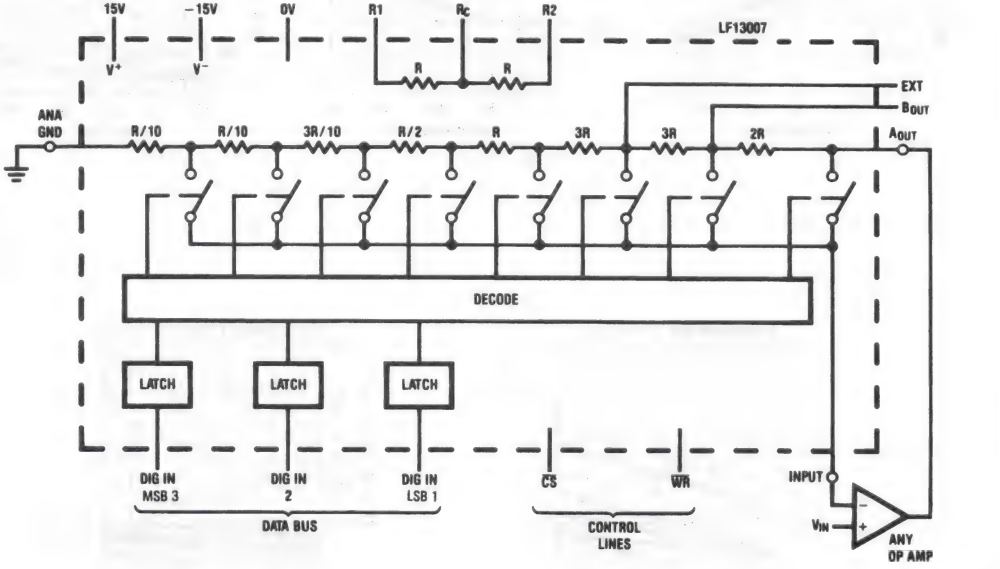
TL/H/5114–2

Switching Waveforms



TL/H/5114–3

Block Diagram and Typical Application (Continued) (LF13007)

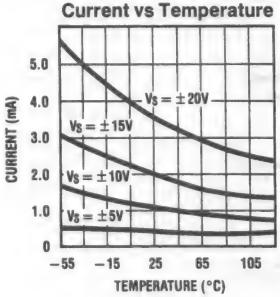


TL/H/5114–4

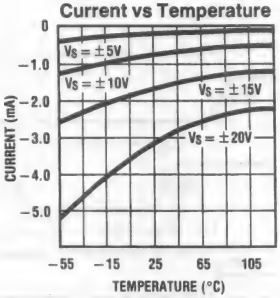
Note: R ≅ 15 kΩ

Typical Performance Characteristics

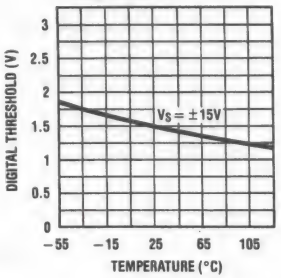
Positive Power Supply Current vs Temperature



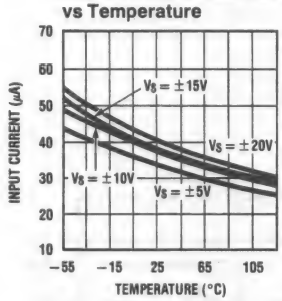
Negative Power Supply Current vs Temperature



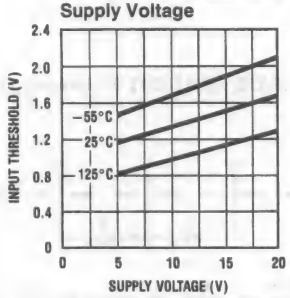
Digital Input Threshold vs Temperature



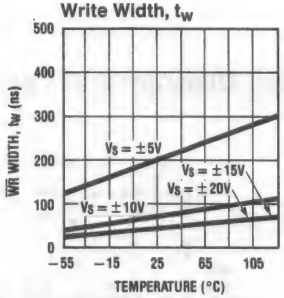
Logical 0 Input Bias Current vs Temperature



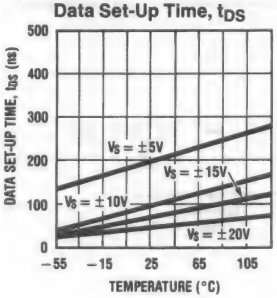
Digital Input Threshold vs Supply Voltage



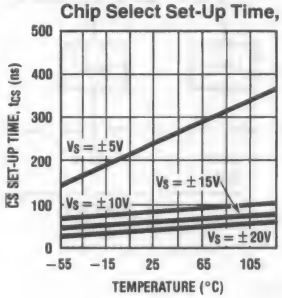
Write Width, t_W



Data Set-Up Time, t_{DS}



Chip Select Set-Up Time, t_{CS}



TL/H/5114–5

Application Information

FLOW-THROUGH OPERATION

THE LF13006, LF13007 can be operated with control lines $\overline{CS}$ and $\overline{WR}$ grounded. In this mode new data on the digital inputs will immediately set the new gain value. Input data cannot be latched in this mode.

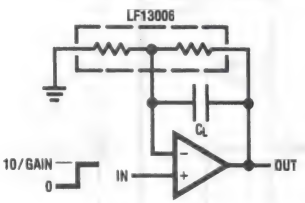
INPUT CURRENT

Current flowing through the input (pin 2) due to bias current of the op amp will result in a gain error due to switch impedance. Normally this error is very small. For example, 10 nA of bias current flowing through 3 kΩ of switch resistance will result in an error of 30 μV at the summing node. However, applications that have significant current flowing through the input must take this effect into account.

SETTLING TIME

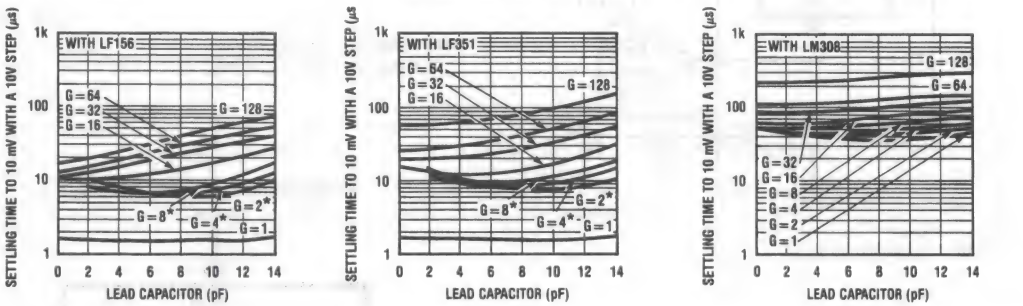
Settling time is a function of the particular op amp used with the LF13006/7 and the gain that is selected. It can be optimized and stability problems can be prevented through the use of a lead capacitor from the inverting input to the output of the amplifier. A lead capacitor is effective whenever the feedback around an amplifier is resistive, whether with discrete resistors or with the LF13006/7. It compensates for the feedback pole created by the parallel resistance and capacitance from the inverting input of the op amp to AC ground.

Settling Time Test Circuit



TL/H/5114–6

Typical Settling Time Curves



TL/H/5114–7

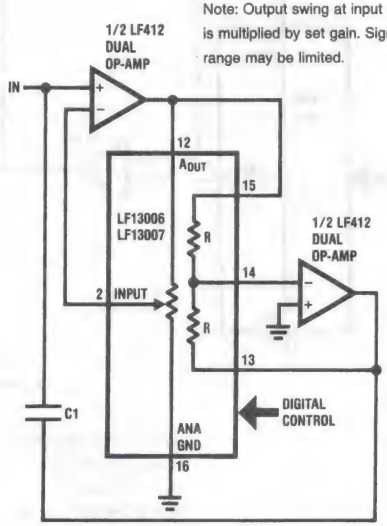
* Unstable at C_L less than 2 pF

Typical Applications

Variable Capacitance Multiplier

$C_{effective} = C1(\text{gain set \#})$

Note: Output swing at input op amp is multiplied by set gain. Signal range may be limited.



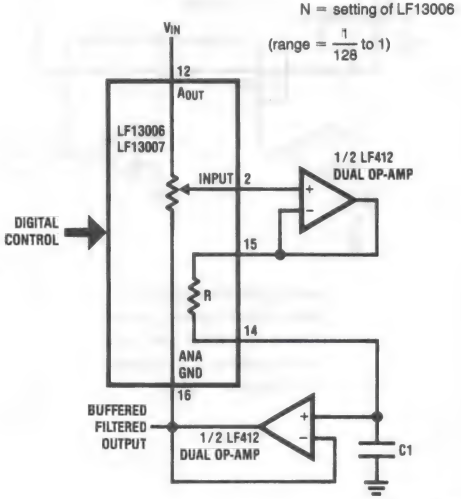
TL/H/5114–8

Variable Time Constant Filter

Time constant = $\frac{R}{N}C1$

N = setting of LF13006

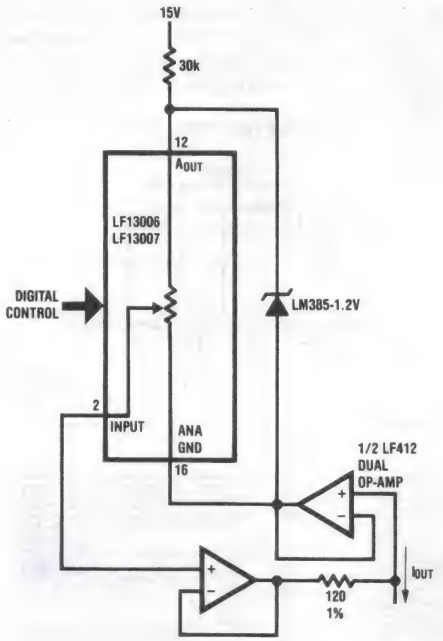
(range = $\frac{1}{128}$ to 1)



TL/H/5114–9

Typical Applications (Continued)

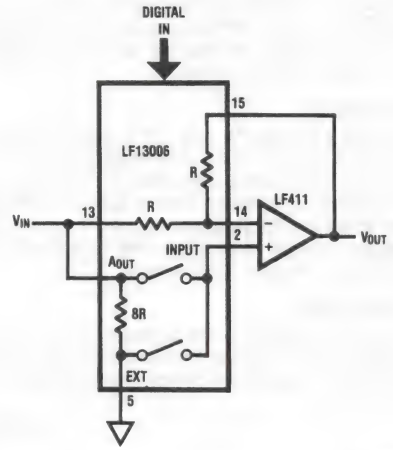
Programmable Current Source



TL/H/5114–10

$$I_{OUT} = \frac{1.2V}{120\Omega}\left[\frac{1}{\text{gain set \#}}\right]$$

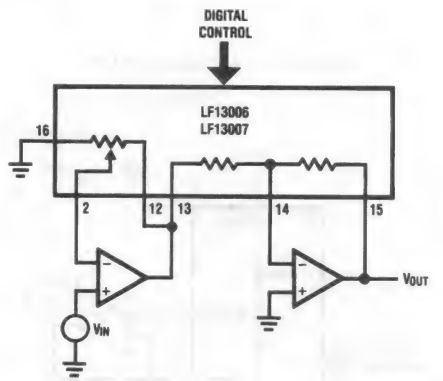
Switchable Gain of ±1



TL/H/5114–11

Note: Digital code = 000, $V_{OUT} = V_{IN}$;
Digital code = 001, $V_{OUT} = -V_{IN}$

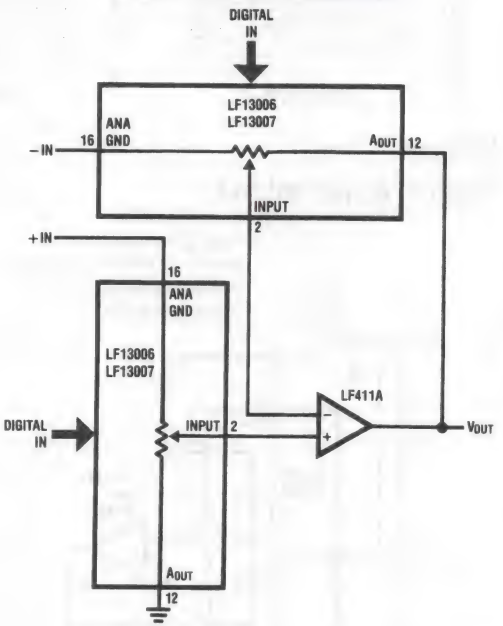
Inverting Gains



TL/H/5114–12

Inverting gain with high input impedance can be obtained with the LF13006, LF13007 by using the two on-board resistors and a dual op amp as shown.

Programmable Differential Amp

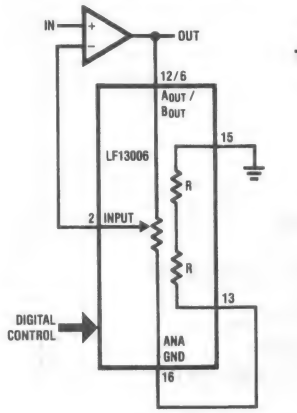


TL/H/5114–13

Note 1: Actual gain = set gain − 1 since LF13006s are in "inverting mode".
Note 2: Set gain must be same on both LF13006s.

Typical Applications (Continued)

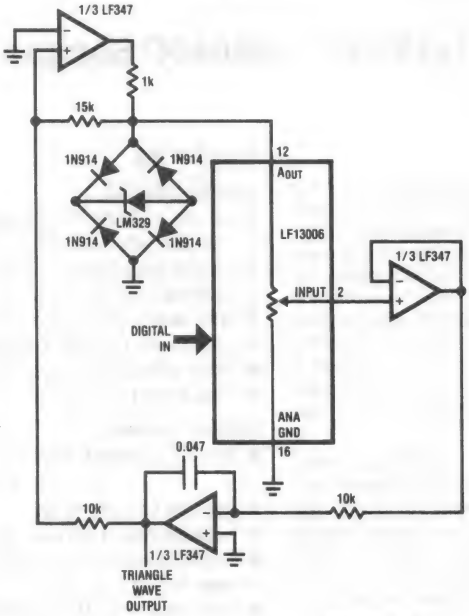
Altered Gain Range



TL/H/5114–14

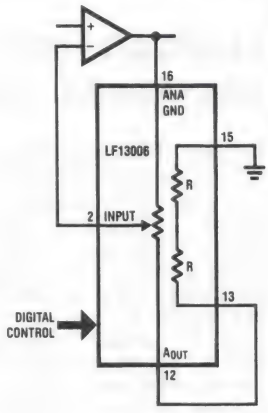
GAINS	
A_{OUT}	B_{OUT}
1	1
1.8	1.2
3	2
4.5	3
6	4
7.2	4.8
8	5.33
8.47	5.65

One Octave per Bit Function Generator



TL/H/5114–15

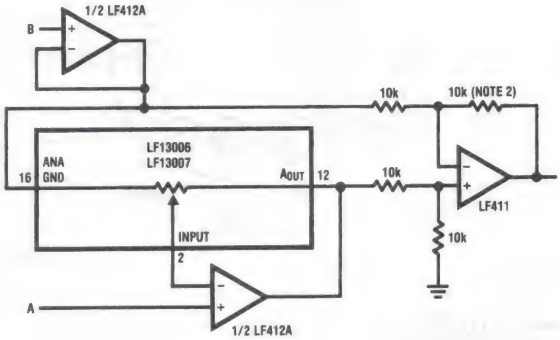
Variable Gains of Almost 1



TL/H/5114–16

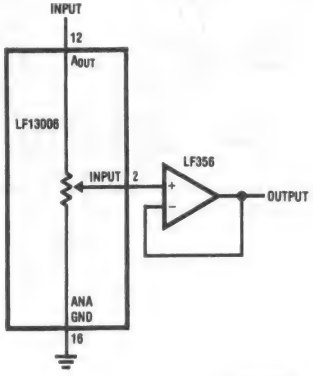
GAINS
9
1.8
1.29
1.125
1.059
1.029
1.014
1.007

Programmable Instrumentation Amp



TL/H/5114–17

Attenuator (0 dB to −42 dB in 6 dB steps)



TL/H/5114–18

Note 1: $V_{OUT} = N (A - B)$, N = set gain.

Note 2: All 10k resistors 0.1% matched.

National Semiconductor

LH0023/LH0023C/LH0043/LH0043C Sample and Hold Circuits

General Description

The LH0023/LH0023C and LH0043/LH0043C are complete sample and hold circuits including input buffer amplifier, FET output amplifier, analog signal sampling gate, TTL compatible logic circuitry and level shifting. They are designed to operate from standard ±15V DC supplies, but provision is made on the LH0023/LH0023C for connection of a separate +5V logic supply in minimum noise applications. The principal difference between the LH0023/LH0023C and the LH0043/LH0043C is a 10:1 trade-off in performance between sample accuracy and sample acquisition time. Devices are pin compatible except for TTL logic polarity.

The LH0023/LH0023C and LH0043/LH0043C are ideally suited for a wide variety of sample and hold applications including data acquisition, analog to digital conversion, synchronous demodulation, and automatic test setup. They offer significant cost and size reduction over equivalent module or discrete designs. Each device is available in a hermetic TO-8 package and is completely specified over both full military and industrial temperature ranges.

The LH0023 and LH0043 are specified for operation over the −55°C to +125°C military temperature range. The LH0023C and LH0043C are specified for operation over the −25°C to +85°C temperature range.

Features

LH0023/LH0023C

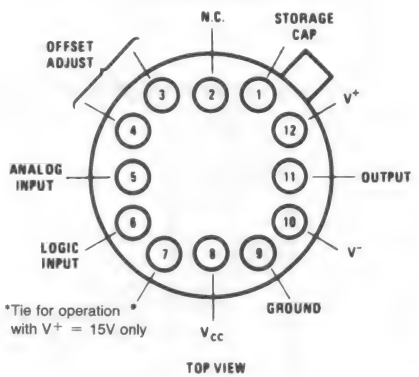
- Sample accuracy–0.01% max
- Hold drift rate–0.5 mV/sec typ
- Sample acquisition time–100 μs max for 20V
- Aperture time–150 ns typ
- Wide analog input range–±10V min
- Logic input–TTL/DTL compatible
- Offset adjustable to zero with single 10k pot
- Output short circuit proof

LH0043/LH0043C

- Sample acquisition time–15 μs max for 20V
  4 μs typ for 5V
- Aperture time–20 ns typ
- Hold drift rate–1 mV/sec typ
- Sample accuracy–0.1% max
- Wide analog input range–±10V min
- Logic input–TTL/DTL compatible
- Offset adjustable to zero with single 10k pot
- Output short circuit protection

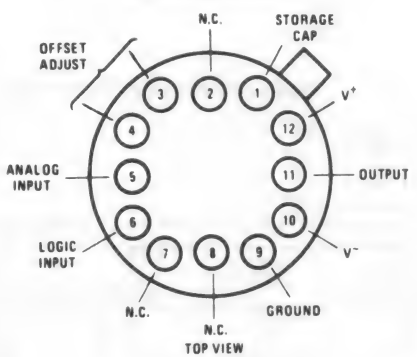
Connection Diagrams

LH0023/LH0023C



TL/K/5693–1

LH0043/LH0043C

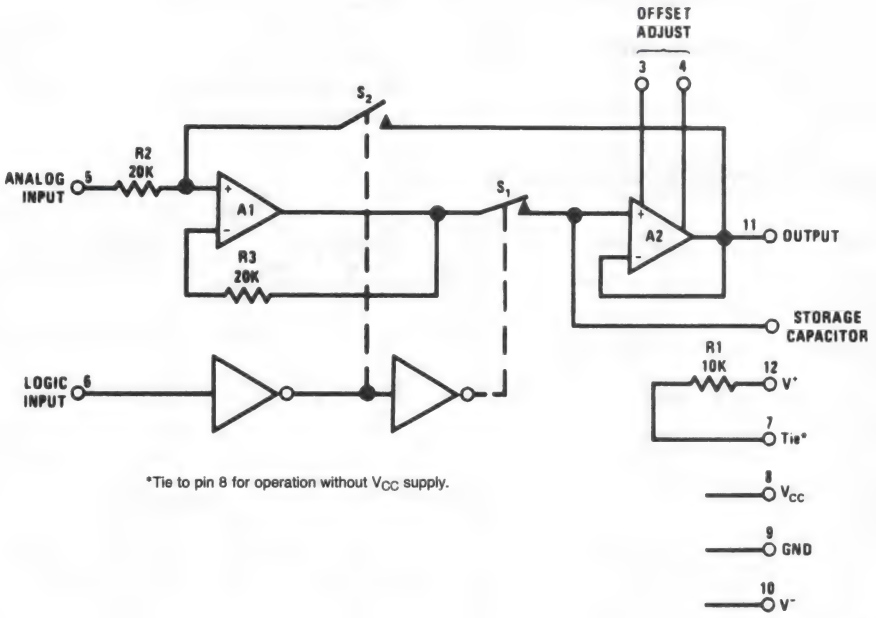


TL/K/5693–8

Order Number LH0023G or LH0023CG or LH0043G or LH0043CG
See Package Number G12B

Block Diagrams

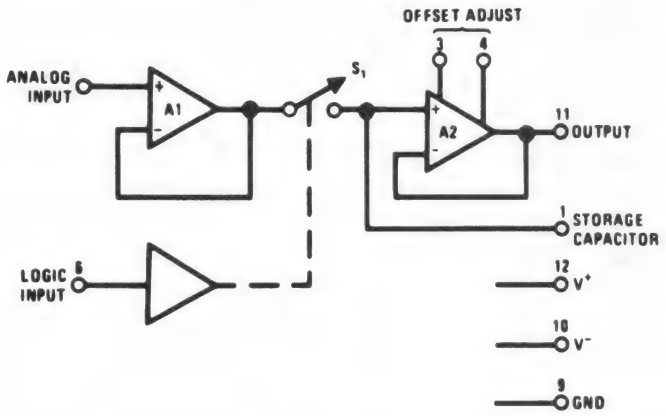
LH0023/LH0023C



*Tie to pin 8 for operation without V_{CC} supply.

TL/K/5693-9

LH0043/LH0043C



TL/K/5693-10

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_S)	±20V
Logic Supply Voltage (V_{CC}) LH0023, LH0023C	+7.0V
Logic Input Voltage (V_6)	+5.5V
Analog Input Voltage (V_5)	±15V
Power Dissipation	See graph
Output Short Circuit Duration	Continuous
Operating Temperature Range	
LH0023, LH0043	−55°C to +125°C
LH0023C, LH0043C	−25°C to +85°C
Storage Temperature Range	−65°C to −150°C
Lead Soldering (10 seconds)	300°C

Electrical Characteristics LH0023/LH0023C (Note 1)

Parameter	Conditions	Limits						Units
		LH0023			LH0023C			
		Min	Typ	Max	Min	Typ	Max	
Sample (Logic "1") Input Voltage	V_{CC} = 4.5V	2.0			2.0			V
Sample (Logic "1") Input Current	V_6 = 2.4V, V_{CC} = 5.5V			5.0			5.0	μA
Hold (Logic "0") Input Voltage	V_{CC} = 4.5V			0.8			0.8	V
Hold (Logic "0") Input Current	V_6 = 0.4V, V_{CC} = 5.5V			0.5			0.5	mA
Analog Input Voltage Range		±10	±11		±10	±11		V
Supply Current – I_{10}	V_5 = 0V, V_6 = 2V, V_{11} = 0V		4.5	6		4.5	6	mA
Supply Current – I_{12}	V_5 = 0V, V_6 = 0.4V, V_{11} = 0V		4.5	6		4.5	6	mA
Supply Current – I_8	V_8 = 5.0V, V_5 = 0		1.0	1.6		1.0	1.6	mA
Sample Accuracy	V_{OUT} = ±10V (Full Scale)		0.002	0.01		0.002	0.02	%
DC Input Resistance	Sample Mode Hold Mode	500 20	1000 25		300 20	1000 25		kΩ kΩ
Input Current – I_5	Sample Mode		0.2	1.0		0.3	1.5	μA
Input Capacitance			3.0			3.0		pF
Leakage Current – pin 1	V_5 = ±10V; V_{11} = ±10V,		10.0	200		20.0	500	pA
	−55°C≤T_A≤125°C V_5 = ±10V; V_{11} = ±10V			5.0			2.0	nA
Drift Rate	V_{OUT} = ±5V, C_S = 0.01 μF, T_A = 25°C		0.5			0.5		mV/s
Drift Rate	V_{OUT} = ±10V, C_S = 0.01 μF, T_A = 25°C		1.0	20		2.0	50	mV/s
Drift Rate	V_{OUT} = ±10V, C_S = 0.01 μF			0.50			0.2	mV/ms
Aperture Time			150			150		ns
Sample Acquisition Time	ΔV_{OUT} = 20V, C_S = 0.01 μF		50	100		50	100	μs
Output Amplifier Slew Rate		1.5	3.0		1.5	3.0		V/μs
Output Offset Voltage (without null)	R_S ≤ 10k, V_5 = 0V, V_6 = 2.0V			±20			±20	mV
Analog Voltage	R_L ≥ 1k, T_A = 25°C	±10	±11		±10	±11		V
Output Range	R_L ≥ 2k	±10	±12		±10	±12		V

Note 1: Unless otherwise noted, these specifications apply for V^+ = +15V, V_{CC} = +5V, V^- = −15V, pin 9 grounded, a 0.01μF capacitor connected between pin 1 and ground over the temperature range −55°C to +125°C for the LH0023, and −25°C to +85°C for the LH0023C. All typical values are for T_A = 25°C.

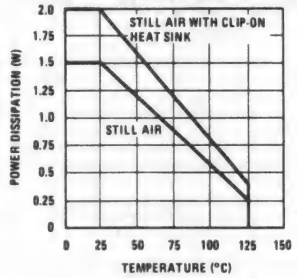
Electrical Characteristics LH0043/LH0043C: (Note 2)

Parameter	Conditions	Limits						Units
		LH0043			LH0043C			
		Min	Typ	Max	Min	Typ	Max	
Hold (Logic "1") Input Voltage		2.0			2.0			V
Hold (Logic "1") Input Current	$V_6 = 2.4V$			5.0			5.0	μA
Sample (Logic "0") Input Voltage				0.8			0.8	V
Sample (Logic "0") Input Current	$V_6 = 0.4V$			1.5			1.5	mA
Analog Input Voltage Range		±10	±11		±10	±11		V
Supply Current	$V_5 = 0V, V_6 = 2V, V_{11} = 0V$ $V_5 = 0V, V_6 = 0.4V,$ $V_{11} = 0V$		20 14	22 18		20 14	22 18	mA mA
Sample Accuracy	$V_{OUT} = \pm 10V$ (Full Scale)		0.02	0.1		0.02	0.3	%
DC Input Resistance	$T_C = 25°C$	10^{10}	10^{12}		10^{10}	10^{12}		Ω
Input Current – I_5			1.0	5.0		2.0	10.0	nA
Input Capacitance			1.5			1.5		pF
Leakage Current– pin 1	$V_5 = \pm 10V; V_{11} = \pm 10,$ $T_C = 25°C$ $V_5 = \pm 10V; V_{11} = \pm 10V$		10 10	25 25		20 2	50 5	pA nA
Drift Rate	$V_{OUT} = \pm 10V, C_S = 0.001\ \mu F,$ $T_C = 25°C$		10	25		20	50	mV/s
Drift Rate	$V_{OUT} = \pm 10V, C_S = 0.001\ \mu F$		10	25		2	5	mV/ms
Drift Rate	$V_{OUT} = \pm 10V, C_S = 0.01\ \mu F,$ $T_C = 25°C$		1	2.5		2	5	mV/s
Drift Rate	$V_{OUT} = \pm 10V, C_S = 0.01\ \mu F$		1	2.5		0.2	0.5	mV/ms
Aperture Time			20	60		20	60	ns
Sample Acquisition Time	$\Delta V_{OUT} = 20V, C_S = 0.001\ \mu F$ $\Delta V_{OUT} = 20V, C_S = 0.01\ \mu F$ $\Delta V_{OUT} = 5V, C_S = 0.001\ \mu F$		10 30 4	15 50		10 30 4	15 50	μs μs μs
Output Amplifier Slew Rate	$V_{OUT} = 5V, C_S = 0.001\ \mu F$	1.5	3.0		1.5	3.0		V/μs
Output Offset Voltage (without null)	$R_S \leq 10k, V_5 = 0V, V_6 = 0V$			±40			±40	mV
Analog Voltage Output Range	$R_L \geq 1k, T_A = 25°C$ $R_L \geq 2k$	±10 ±10	±11 ±12		±10 ±10	±11 ±12		V V

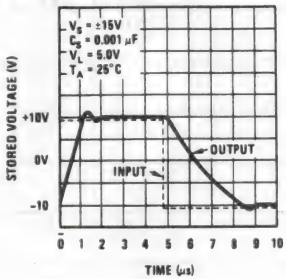
Note 2: Unless otherwise noted, these specifications apply for $V^+ = \pm 15V$, $V^- = -15V$, pin 9 grounded, a 5000 pF capacitor connected between pin 1 and ground over the temperature range −55°C to +125°C for the LH0043, and −25°C to +85°C for the LH0043C. All typical values are for $T_C = 25°C$.

Typical Performance Characteristics

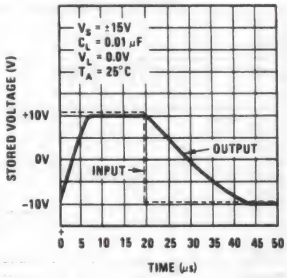
Power Dissipation



Sample Acquisition Time—LH0043



Sample Acquisition Time—LH0043



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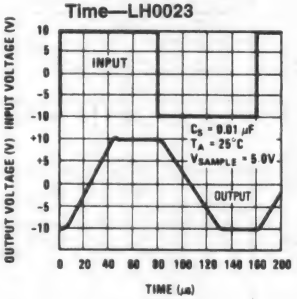
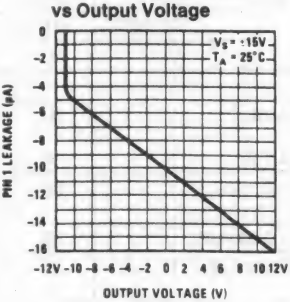
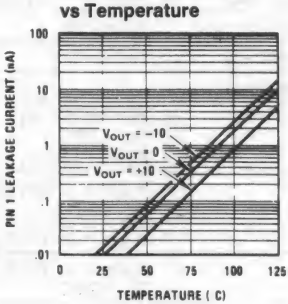
Sample Acquisition Time—LH0023



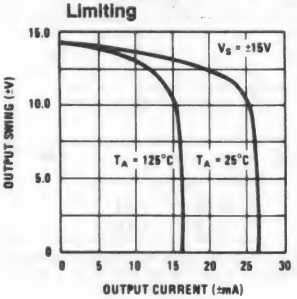
Pin 1 Leakage Current vs Output Voltage



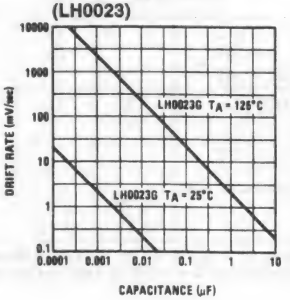
Pin 1 Leakage Current vs Temperature



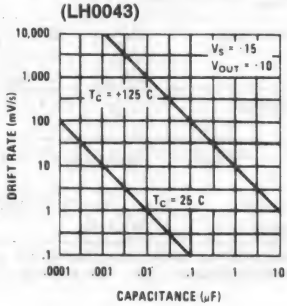
Output Current Limiting



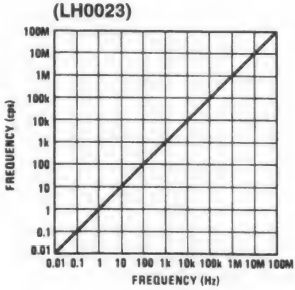
Drift vs Capacitance (LH0023)



Drift vs Capacitance (LH0043)



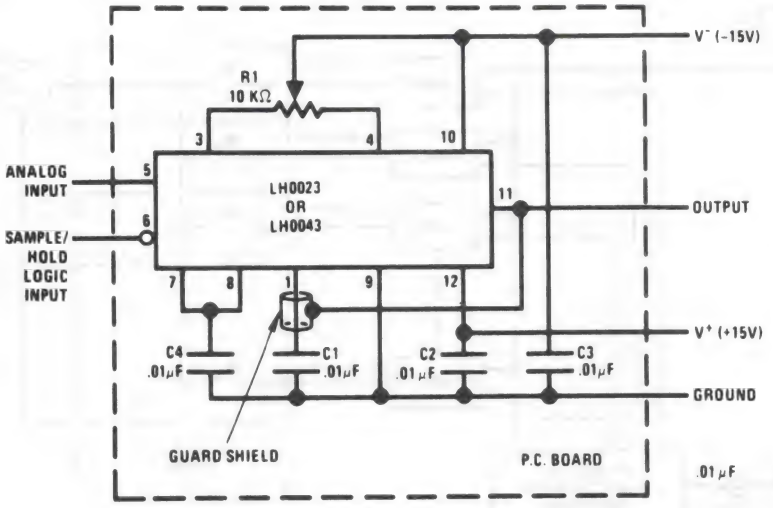
(LH0023)



TL/K/5693-3

Typical Applications

How to Build a Sample and Hold Module



TL/K/5693-4

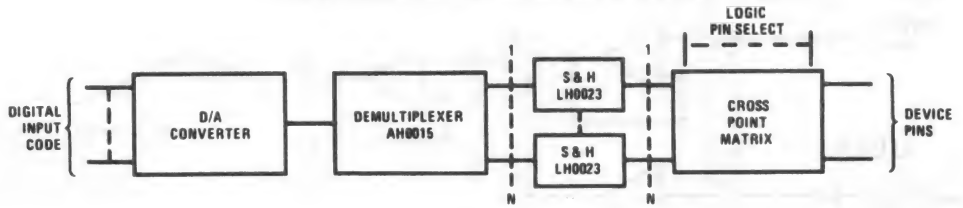
Note 1: C1 is polystyrene.

Note 2: C2, C3, C4 are ceramic disc.

Note 3: Jumper 7–8 and C4 not required for LH0043.

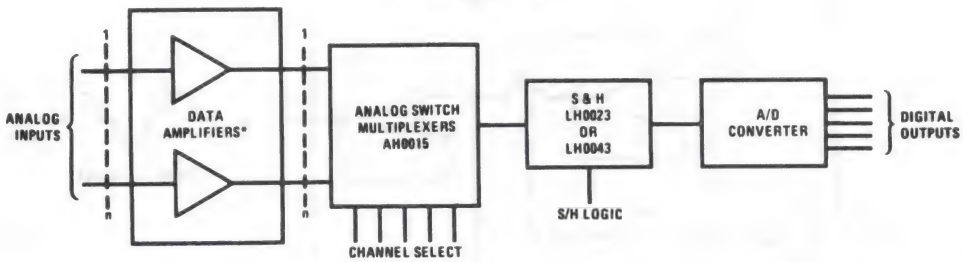
Note 4: R1 optional if zero trim is required.

Forcing Function Setup for Automatic: Test Gear



TL/K/5693-5

Data Acquisition System

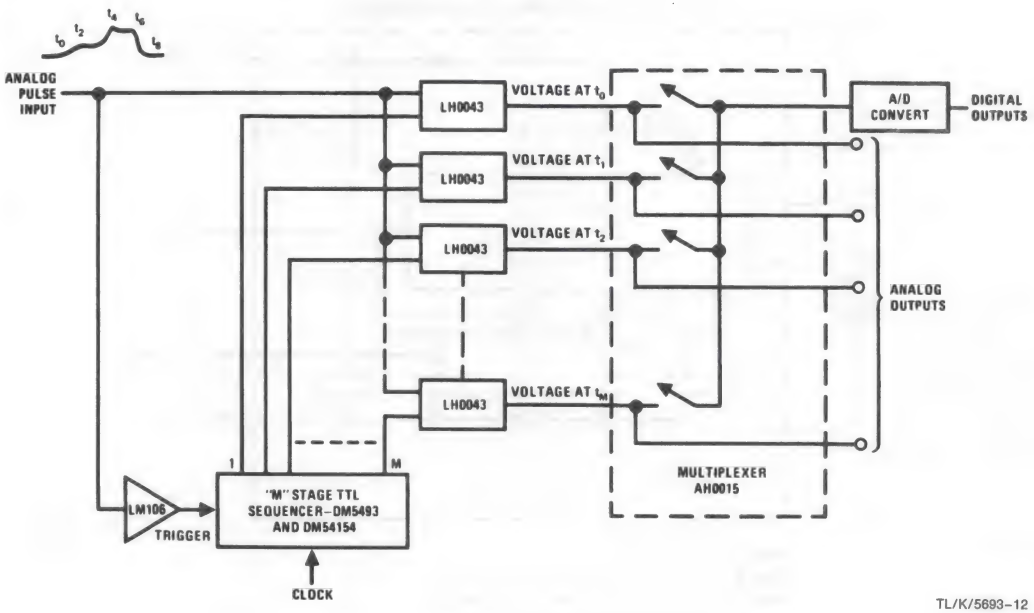


TL/K/5693-11

*See op amp selection guide for details. Most popular types include LH0052, LM108, LM112, LH0044, LH0036, and LH0038.

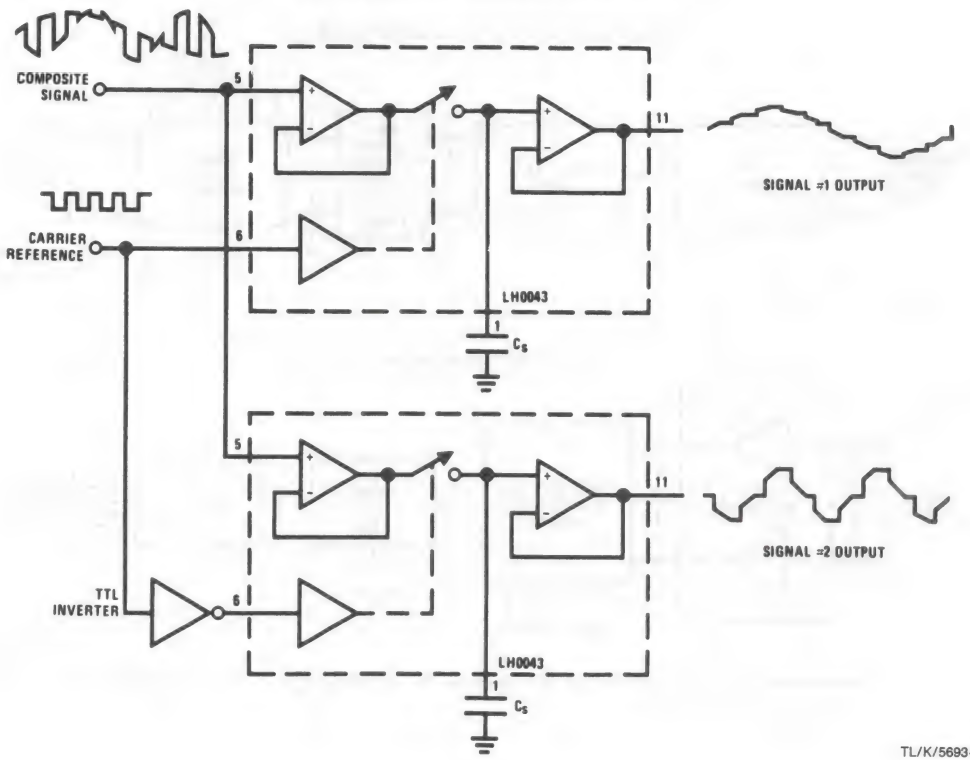
Typical Applications (Continued)

Single Pulse Sampler



TL/K/5693–12

Two Channel Double Sideband Demodulator



TL/K/5693–13

Applications Information

1.0 DRIFT ERROR MINIMIZATION

In order to minimize drift error, care in selection of C_S and layout of the printed circuit board is required. The capacitor should be of high quality Teflon, polycarbonate, or polystyrene construction. Board cleaniness and layout are critical particularly at elevated temperatures. See AN-63 for detailed recommendations. A guard conductor connected to the output surrounding the storage node (pin 1) will be helpful in meeting severe environmental conditions which would otherwise cause leakage across the printed circuit board.

2.0 CAPACITOR SELECTION

The size of the capacitor is dictated by the required drift rate and acquisition time. The drift is determined by the leakage current at pin 1 and may be calculated by $\frac{dV}{dt} = \frac{I_L}{C_s}$, where I_L is the total leakage current at pin 1 of the device, and C_s is the value of the storage capacitor.

2.1 Capacitor Selection – LH0023

At room temperature leakage current for the LH0023 is approximately 100 pA. A drift rate of 10 mV/sec would require a 0.01 μF capacitor.

For values of C_s up to 0.01 μF the acquisition time is limited by the slew rate of the input buffer amplifier, A1, typically 0.5 V/μs. Beyond this point, current availability to charge C_s also enters the picture. The acquisition time is given by:

$$t_A \cong \sqrt{\frac{2\Delta e_O RC_s}{0.5 \times 10^6}} = 2 \times 10^{-3}\sqrt{\Delta e_O RC_s}$$

where: R = the internal resistance in series with C_s

Δe_o = change in voltage sampled

An average value for R is approximately 600 ohms. The expression for t_A reduces to:

$$t_A \cong \frac{\sqrt{\Delta e_o C_s}}{20}$$

For a −10V to +10V change and C_s = .05 μF, acquisition time is typically 50 μs.

2.2 Capacitor Selection–LH0043

At 25°C case temperature, the leakage current for the LH0043G is approximately 10 pA, so a drift rate of 5 mV/s would require a capacitor of $C_s = 10 \bullet 10^{12}/5 \bullet 10^3 = 2000$ pF or larger.

For values of C_s below about 5000 pF, the acquisition time of the LH0043G will be limited by the slew rate of the output amplifier (the signal will be acquired, in the sense that the voltage will be stored on the capacitor, in much less time as dictated by the slew rate and current capacity of the input amplifier, but it will not be available at the output). For larger values of storage capacitance, the limitation is the current sinking capability of the input amplifier, typically 10 mA. With $C_S = 0.01$ μF, the slew rate can be estimated by

$$\frac{dV}{dt} = \frac{10 \bullet 10^{-3}}{0.01 \bullet 10^{-6}} = 1V/\mu s$$ or a slewing time for a 5 volt signal change of 5μs.

3.0 OFFSET NULL

Provision is made to null both the LH0023 and LH0043 by use of a 10k pot between pins 3 and 4. Offset null should be accomplished in the sample mode at one half the input voltage range for minimum average error.

4.0 SWITCHING SPIKE MINIMIZATION–LH0043

A capacitive divider is formed by the storage capacitor and the capacitance of the internal FET switch which causes a small error current to be injected into the storage capacitor at the termination of the sample interval. This can be considered a negative DC offset and nulled out as described in (3.0), or the transient may be nulled by coupling an equal but opposite signal to the storage capacitor. This may be accomplished by connecting a capacitor of about 30 pF (or a trimmer) between the logic input (pin 6) and the storage capacitor (pin 1). Note that this capacitor must be chosen as carefully as the storage capacitor itself with respect to leakage. The LH0023 has switch spike minimization circuitry built into the device.

5.0 ELIMINATION OF THE 5V LOGIC SUPPLY – LH0023

The 5V logic supply may be eliminated by shorting pin 7 to pin 8 which connects a 10k dropping resistor between the +15V and V_C. Decoupling pin 8 to ground through 0.1 μF disc capacitor is recommended in order to minimize transients in the output.

6.0 HEAT SINKING

The LH0023 and LH0043G may be operated without damage throughout the military temperature range of −55 to +125°C (−25 to +85°C for the LH0023CG and LH0043CG) with no explicit heat sink, however power dissipation will cause the internal temperature to rise above ambient. A simple clip-on heat sink such as Wakefield #215−1.9 or equivalent will reduce the internal temperature about 20°C thereby cutting the leakage current and drift rate by one fourth at max. ambient. There is no internal electrical connection to the case, so it may be mounted directly to a grounded heat sink.

7.0 THEORY OF OPERATION–LH0023

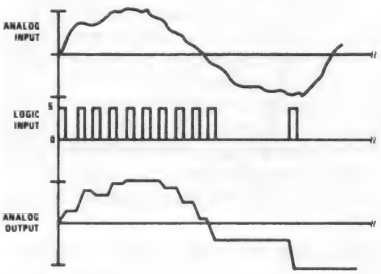
The LH0023/LH0023C is comprised of input buffer amplifier, A1, analog switches, S1 and S2, a TTL to MOS level

Applications Information (Continued)

translator, and output buffer amplifier, A2. In the "sample" mode, the logic input is raised to logic "1" ($V_6 \leq 2.0V$) which closes S1 and opens S2. Storage capacitor, C_s, is charged to the input voltage through S1 and the output slews to the input voltage. In the "hold" mode, the logic input is lowered to logic "0" ($V_6 \leq 0.8V$) opening S1 and closing S2. C_s retains the sample voltage which is applied to the output via A2. Since S1 is open, the input signal is overriden, and leakage across the MOS switch is therefore minimized. With S1 open, drift is primarily determined by input bias current of A2, typically 100 pA at 25°C.

7.1 Theory of Operation–LH0043

The LH0043/LH0043C is comprised of input buffer amplifier A1, FET switch S1 operated by a TTL compatible level translator, and output buffer amplifier A2. To enter the "sample" mode, the logic input is taken to the TTL logic "0" state ($V_6 = 0.8V$) which commands the switch S1 closed and allows A1 to make the storage capacitor voltage equal to the analog input voltage. in the "hold" mode ($V_6 = 2.0V$), S1 is opened isolating the storage capacitor from the input and leaving it charged to a voltage equal to the last analog input voltage before entering the hold mode. The storage capacitor voltage is brought to the output by low leakage amplifier A2.



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8.0 DEFINITIONS

V_5: The voltage at pin 5, e.g., the analog input voltage.

V_6: The voltage at pin 6, e.g., the logic control input signal.

V_{11}: The voltage at pin 11, e.g., the output signal.

T_A: The temperature of the ambient air.

T_C: The temperature of the device case at the center of the bottom of the header.

Acquisition Time:

The time required for the output (pin 11) to settle within the rated accuracy after a specified input change is applied to the input (pin 5) with the logic input (pin 6) in the low state.

Aperture Time:

The time indeterminacy when switching from sample mode to hold including the delay from the time the mode control signal (pin 6) passes through its threshold (1.4 volts) to the time the circuit actually enters the hold mode.

Output Offset Voltage:

The voltage at the output terminal (pin 11) with the analog input (pin 5) at ground and logic input (pin 6) in the "sample" mode. This will always be adjustable to zero using a 10k pot between pins 3 and 4 with the wiper arm returned to V^-.

National Semiconductor

LH0053/LH0053C High Speed Sample and Hold Amplifier

General Description

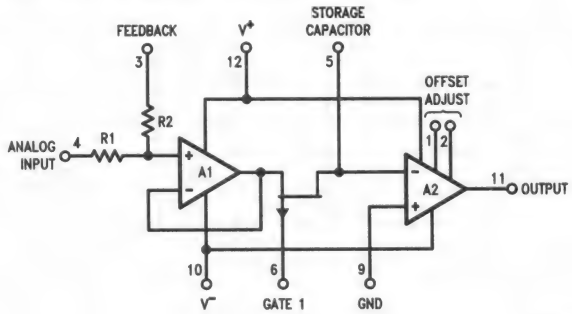
The LH0053/LH0053C is a high speed sample and hold circuit capable of acquiring a 20V step signal in under 5.0 μs.

The device is ideally suited for a variety of high speed data acquisition applications including analog buffer memories for A to D conversion and synchronous demodulation.

Features

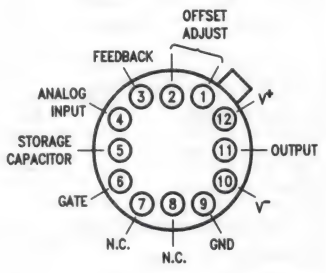
- Sample acquisition time 10 μs max. for 20V signal
- FET switch for preset or reset function
- Sample accuracy null
- Offset adjust to 0V
- DTL/TTL compatible FET gate
- Single storage capacitor

Schematic and Connection Diagrams



TL/H/9251-1

Metal Can Package



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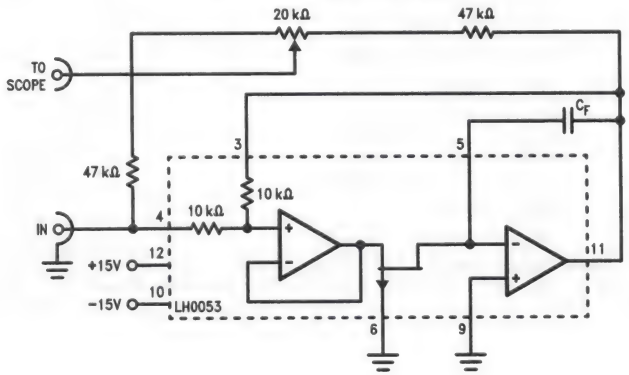
Top View

Order Number LH0053G or LH0053CG
See NS Package Number G12B

AC Test Circuit

Acquisition Time Test Circuit



TL/H/9251-3

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V^+ and V^-)	±18V
Gate Input Voltage (V_6)	±20V
Analog Input Voltage (V_4)	±15V
Input Current (I_8 and I_5)	±10 mA
Power Dissipation	See graph
Output Short Circuit Duration	Continuous
Operating Temperature Range	
LH0053	−55°C to +125°C
LH0053C	−25°C to +85°C
Storage Temperature Range	−65°C to +150°C
Lead Temperature (Soldering, 10 sec.)	300°C

Electrical Characteristics (Note 1)

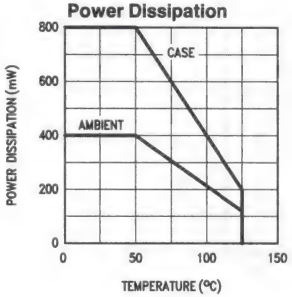
Parameter	Conditions	Limits						Units
		LH0053			LH0053C			
		Min	Typ	Max	Min	Typ	Max	
Sample (Gate "0") Input Voltage				0.5			0.5	V
Sample (Gate "0") Input Current	V_6 = 0.5V, T_A = 25°C V_6 = 0.5V			−5.0 −100			−5.0 −100	µA µA
Hold (Gate "1") Input Voltage		4.5			4.5			V
Hold (Gate "1") Input Current	V_6 = 4.5V, T_A = 25°C V_6 = 4.5V			1.0 1.0				nA µA
Analog Input Voltage Range		±10	±11		±10	±11		V
Supply Current	V_4 = 0V, V_6 = 0.5V		13	18		13	18	mA
Input Bias Current (I_4)	V_4 = 0V, T_A = 25°C		120	250		150	500	nA
Input Resistance		5.0	10	15	5.0	10	15	kΩ
Analog Output Voltage Range	R_L = 2.0 kΩ	±10	±12		±10	±12		V
Output Offset Voltage	V_4 = 0V, V_6 = 0.5V, T_A = 25°C V_4 = 0V, V_6 = 0.5V		5.0	7.0 10		5.0	10 15	mV mV
Sample Accuracy (Note 2)	V_4 = ±10V, V_6 = 0.5V, T_A = 25°C		0.1	0.2		0.1	0.3	%
Aperture Time	ΔV_6 = 4.5V, T_A = 25°C		10	25		10	25	ns
Sample Acquisition Time	V_4 = ±10V, T_A = 25°C, C_F = 1000 pF, V_6 = 0V		5.0	10		8.0	15	µs
Sample Acquisition Time	V_4 = ±10V, T_A = 25°C, C_F = 100 pF, V_6 = 0V		4.0			4.0		µs
Output Slew Rate	ΔV_{IN} = ±10V, T_A = 25°C, C_F = 100 pF, V_6 = 0V		20			20		V/µs
Large Signal Bandwidth	V_4 = ±10V, T_A = 25°C, C_F = 1000 pF		200			200		kHz
Large Current (Pin 5)	V_4 = ±10V, T_A = 25°C, V_4 = ±10V		6.0	50 30		10	100 30	pA nA
Drift Rate	V_4 = ±10V, T_A = 25°C, C_F = 1000 pF		6.0	50		10	100	mV/s
Drift Rate	V_4 = ±10V, C_F = 1000 pF			30			30	V/s

Note 1: Unless otherwise noted, these specifications apply for V_S = ±15V, pin 9 grounded, a 1000 pF capacitor between pin 5 and pin 11, pin 3 shorted to pin 11, over the temperature range −55°C to +125°C for the LH0053 and −25°C to +85°C for the LH0053C. All typical values are for T_A = 25°C.

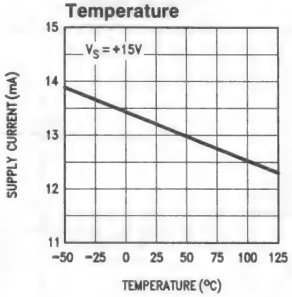
Note 2: Sample accuracy may be nulled by inserting a potentiometer in the feedback loop. This compensates for source impedance and feedback resistor tolerances.

Typical Performance Characteristics

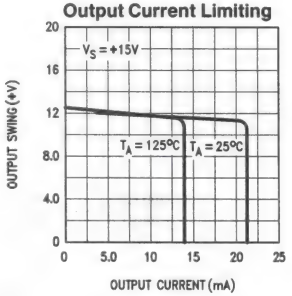
Power Dissipation



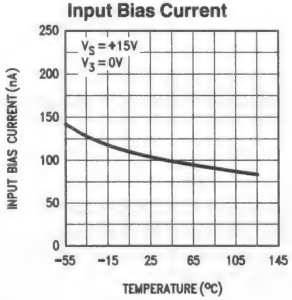
Supply Current vs Temperature



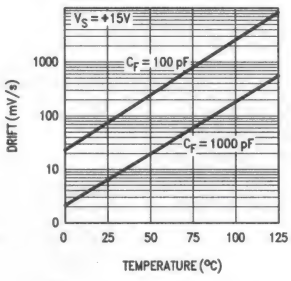
Output Current Limiting



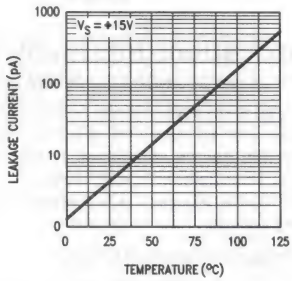
Input Bias Current



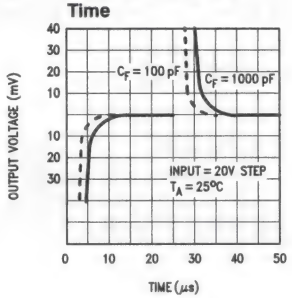
Drift



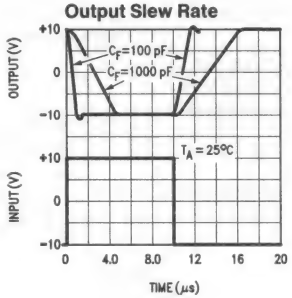
Leakage Current at Pin 5



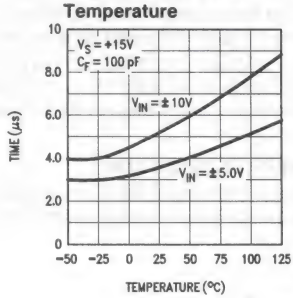
Sample Acquisition Time



Output Slew Rate



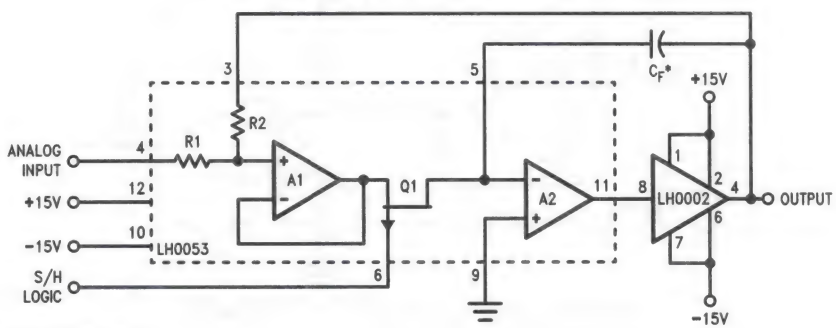
Acquisition Time vs Temperature



TL/H/9251–4

Typical Applications

Increasing Output Drive Capability

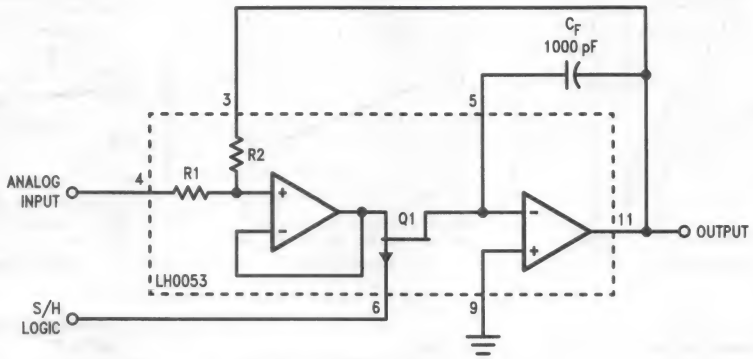


*Polystyrene construction.

TL/H/9251–5

Typical Applications (Continued)

Sample and Hold



TL/H/9251–6

Applications Information

SOURCE IMPEDANCE COMPENSATION

The gain accuracy (linearity) of the LH0053/LH0053C is set by two internal precision resistors. Circuit applications in which the source impedance is non-zero will result in a closed loop gain error, e.g. if $R_S = 10\Omega$, a gain error of 0.1% results. *Figures 1* and *2* show methods for accommodating non-zero source impedance.

DRIFT ERROR MINIMIZATION

In order to minimize drift error, care in selecting C_F and layout of the printed circuit board are required. The capacitor should be of high quality teflon, polycarbonate, or polystyrene construction. Board layout and clean lines are critical particularly at elevated temperature.

A ground guard (shield) surrounding pin 5 will minimize leakage currents to and from the summing junction, arising from extraneous signals. See AN-63 for detailed recommendations.

CAPACITOR SELECTION

The size of the capacitor is determined by the required drift rate usually at the expense of acquisition time.

The drift is dictated by leakage current at pin 5 and is given by:

$$\frac{dv}{dt} = \frac{I_L}{C_F}$$

Where I_L is the leakage current at pin 5 and C_F is the value of the capacitance. The room temperature leakage of the LH0053 is typically 6.0 pA, and a 1000 pF capacitor will yield a drift rate of 6.0 mV per second.

For values of C_F below 1000 pF, acquisition for the LH0053 is primarily governed by the slew rate of the input amplifier (20 V/μs) and the setting time of the output amplifier ($\cong$1.0 μs). For values above C_F = 1000 pF, acquisition time is given by:

$$t_a = \frac{C_F\,\Delta V}{I_{DSS}} + t_{S2}$$

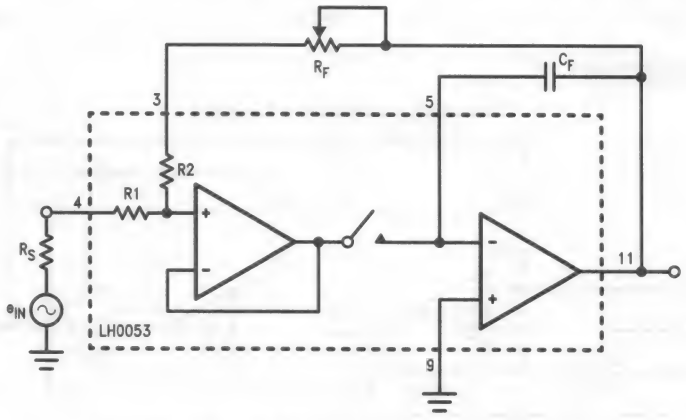
Where:

C_F = The value of the capacitor

ΔV = The magnitude of the input step, e.g. 20V

I_{DSS} = The ON current of switch Q1 $\cong$ 5.0 mA

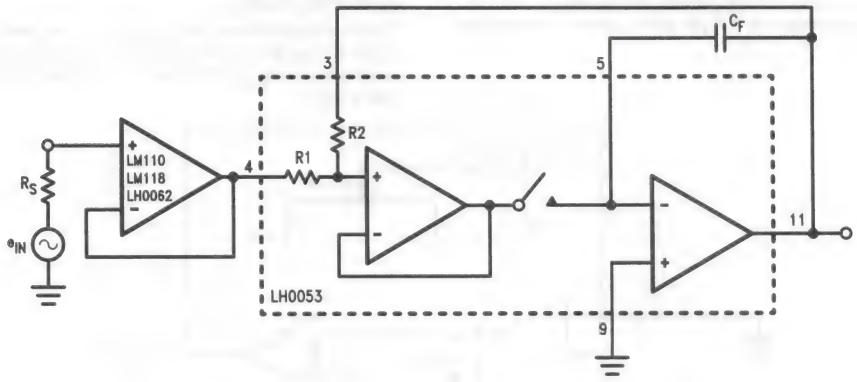
t_{S2} = The setting time of output amplifier $\cong$ 1.0 μs



TL/H/9251–7

FIGURE 1. Non-Zero Source Impedance Compensation

Applications Information (Continued)



TL/H/9251–8

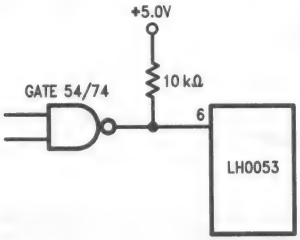
FIGURE 2. Non-Zero Source Impedance Buffering

GATE INPUT CONSIDERATIONS

5.0V TTL Applications

The LH0053 Gate input (pin 6) will interface directly with 5.0V TTL. However, TTL gates typically pull up to 2.5V in the logic "1" state. It is therefore advisable to use a 10 kΩ pull-up resistor between the 5.0V, V_{CC}, and the output of the gate as shown in *Figure 3*.

To obtain the highest speed and fastest acquisition time, the gate drive shown in *Figure 6* is recommended.



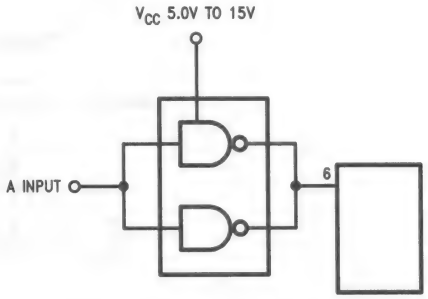
TL/H/9251–9

FIGURE 3. TTL Logic Compatibility

CMOS APPLICATIONS

The LH0053 gate input may be interfaced directly with 74C, CMOS operating off of V_{CC}'s from 5.0V to 15V. However, transient currents of several milliamps can flow on the rising and falling edges of the input signal. It is, therefore, advisable to parallel the outputs of two 54C/74C gates as shown in *Figure 4*.

It should be noted that leakage at pin 5 in the hold mode will be increased by a factor of 2 to 3 when operating into 15V logic levels.



TL/H/9251–10

FIGURE 4. CMOS Logic Compatibility

HEAT SINKING

The LH0053 may be operated over the military temperature range, −55°C to +125°C, without incurring damage to the device. However, a clip on heat sink such as the Wakefield 215 Series or Thermalloy 2240 will reduce the internal temperature rise by about 20°C. The result is two-fold improvement in drift rate at an ambient temperature of 25°C.

Since the case of the device is electrically isolated from the circuit, the LH0053 may be mounted directly to a grounded heat sink.

POWER SUPPLY DECOUPLING

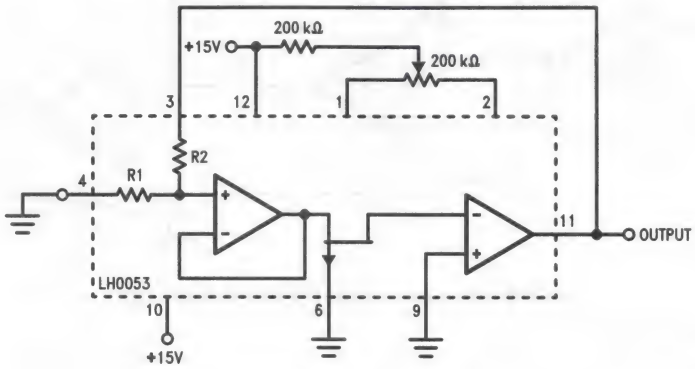
Amplifiers A1 and A2 within the LH0053 are very wide band devices and are sensitive to power supply inductance. It is advisable to bypass V^+ (pin 12) and V^- (pin 10) to ground with 0.1 μF disc capacitors in order to prevent oscillation.

Applications Information (Continued)

Should this procedure prove inadequate, the disc capacitors should be paralleled with 4.7 μF solid tantalum electrolytic capacitors.

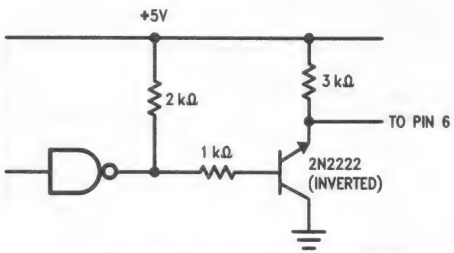
DC OFFSET ADJUST

Output offset error may be adjusted to zero using the circuit shown in *Figure 5*. Offset null should be accomplished in the sample mode ($V_6 \leq 0.5V$) and analog input (pin 4) equal to zero volts.



TL/H/9251–11

FIGURE 5. Offset Null Circuit



TL/H/9251–12

FIGURE 6. High Speed Gate Drive Circuit

Definition of Terms

Voltage V_4: The voltage at pin 4, i.e., the analog input voltage.

Voltage V_6: The voltage at pin 6, i.e., the logic control signal. A logic "1" input, $V_6 \leq 4.5V$, places the LH0053 in the HOLD mode; a logic "0" input ($V_6 \leq 0.5V$) places the device in sample mode.

Acquisition Time: The time required for the output (pin 11) to settle within the rated accuracy after a specified input change is applied to Analog Input 1 (pin 4) with logic input, (pin 6) in the logic "0" state.

Aperture Time: The time indeterminacy when switching from the "sample" mode to the HOLD mode measured from the time the logic input passes through its threshold (2.0V) to the time the device actually enters the HOLD mode.

Sample Accuracy: Difference between input voltage and output voltage while in the sample mode, expressed as a percent of the input voltage.

National Semiconductor

LH4860
Super Fast 12-Bit Track-Hold Amplifier

General Description

The LH4860 is an extremely fast high resolution Sample-Hold (track-and-hold) amplifier. It guarantees acquisition time and sample-to-hold settling time to ±0.01%. The LH4860 will acquire a full 10V signal to ±0.01% full scale (or ±1 mV) in less than 200 ns. The bandwidth of the tracking amplifier is 16 MHz. In the track mode, offset error is typically ±0.5 mV and gain error is typically ±0.05%. The LH4860 is precisely laser trimmed for pedestal compensation. The "Hold" capacitor is internal for ease of use. Also, the bypassing power supply capacitors are inside the package.

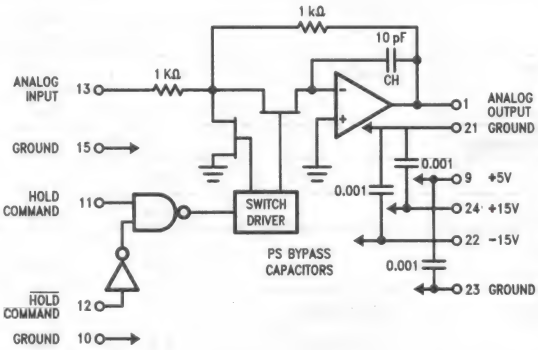
Features

- 200 ns max acquisition time 10V step to ±0.01% FS
- 100 ns max sample-to-hold settling time
- ±50 ps aperture jitter
- 74 dB feedthrough attenuation
- TTL compatible
- Direct replacement for HTC-0300, 4860, and HS9720

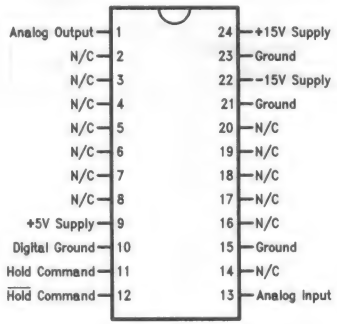
Applications

- Transient recorders
- Fast fourier analysis
- High speed DAS's
- High speed DDS's
- Analog delay and storage

Block and Connection Diagrams



TL/K/9770–2



TL/K/9770–1

Top View

Dual-In-Line Metal Package (D)
Order Number LH4860D or LH4860CD
See NS Package Number D24I

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (V_S^+ and V_S^-)	±18V
Logic Supply Voltage (V_D)	+7V
Analog Input Voltage	$\pm V_S$
Digital Input Voltage	−0.5V to +5.5V
Output Current (Note 1)	±65 mA
Output Short Circuit Duration	Continuous
Operating Temperature Range	
LH4860C	−25°C to +85°C
LH4860	−55°C to +125°C
Storage Temperature Range	−65°C to +150°C
Power Dissipation (P_D) (See Graph)	2.4W
ESD (Note 6)	TBD

DC Electrical Characteristics $V_S = \pm 15V$ and $+5V$, $T_A = +25°C$, unless otherwise noted (Note 1)

Symbol	Parameter	Conditions	LH4860C/LH4860			Units (Max Unless Otherwise Stated)
			Typ	Tested Limit (Note 8)	Design Limit (Note 9)	
	Input/Output Voltage Range		±11.5	±10.25		V (Min)
	Input Impedance		1			kΩ
	Output Current	(Note 1)		40		mA
	Output Impedance		0.1			Ω
	Maximum Capacitive Load		150			pF
	Logic High "1"	(Note 2)		2.0		V (Min)
				5.0		V
	Logic Low "0"	(Note 2)		0		V (Min)
				0.8		V
	Digital Input Loading		1			TTL Load
	Gain		−1.00			V/V
	Gain Accuracy		±0.05	±0.2		%
	Gain Linearity Error (Note 4)		±0.003	±0.01		% FS
	Offset Voltage	Sample Mode	±0.5	±5		mV
	Hold Step	Pedestal *Figure 1*	±2.5	±20		mV
	Gain Drift	(Note 7)	±0.5		**±5**	ppm of FSR/°C
	Offset Drift	Sample Mode (Note 4)	±3		**±15**	ppm of FSR/°C

AC Electrical Characteristics $V_S = \pm 15V$ and $+5V$, $T_A = +25°C$, unless otherwise noted (Note 1)

Symbol	Parameter	Conditions	LH4860C/LH4860			Units (Max Unless Otherwise Stated)
			Typ	Tested Limit (Note 8)	Design Limit (Note 9)	
	Acquisition Time (Notes 4, 5)	10V Step to ±0.01% FS (±1 mV)	150	200		ns
		10V Step to ±0.1% FS (±10 mV)	100	170		ns
		10V Step to ±1% FS (±100 mV)	90			ns
		1V Step to ±1% FS (±100 mV)	75			ns
	Settling Time Sample to Hold (Note 4)	to ±0.01% FS (1 mV)	60	100		ns
		to ±0.1% FS (10 mV)	40			ns
	Sample to Hold Transient		180			mV_{P-P}
	Aperture Delay Time		6			ns
	Aperture Jitter		±50			ps
	Output Slew Rate		300			V/μs
	Small Signal Bandwidth (−3 dB)		16			MHz
	Droop Rate		±0.5	±5		μV/μs
		+85°C	±55			μV/μs
		+125°C	±1.2			mV/μs
	Feedthrough	2.5 MHz, 20 V_{P-P} Input	74			dB
PSRR	Power Supply Rejection Ratio		±0.5			mV/V
	Quiescent Current Drain	+15V Supply	21	25		mA
		−15V Supply	−22	−25		mA
		+5V Supply	17	25		mA
	Power Consumption		730	875		mW

Note 1: The LH4860 output is current limited at approximately ±65 mA and the unit can withstand a sustained short-to-ground. For normal operation, load current should not exceed ±40 mA.

Note 2: See Application Information for use of Hold and $\overline{\text{Hold}}$ inputs.

Note 3: The Hold Command inputs appear as one TTL load and are defined as sinking 40 μA with logic "1" applied and sourcing 1.6 mA with logic "0" applied.

Note 4: FS means "Full Scale" and is equivalent to 10V. FSR means "Full Scale Range" and is equivalent to 20V. For a 12-bit system, 1 LSB = 0.024% FS.

Note 5: Acquisition time is tested with no load.

Note 6: The test circuit used consists of the human body model of 100 pF in series with 1500Ω.

Note 7: **Boldface** limits are guaranteed over full temperature range. Operating ambient temperature range of LH4860C is −25°C to +85°C, and LH4860 is −55°C to +125°C.

Note 8: Tested limits are guaranteed and 100% production tested.

Note 9: Design limits are guaranteed (but not production tested) over the indicated temperature or temperature range. These limits are not used to calculate outgoing quality level.

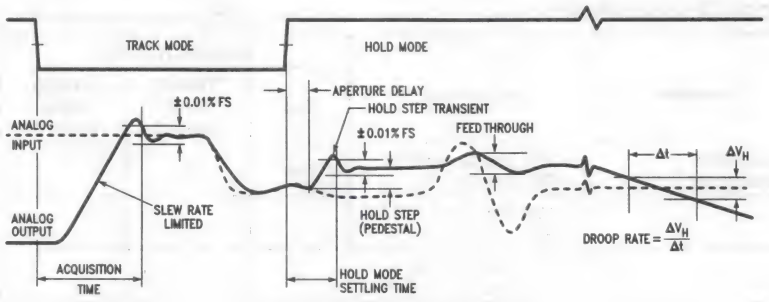


FIGURE 1. Timing Diagram

Typical Performance Characteristics

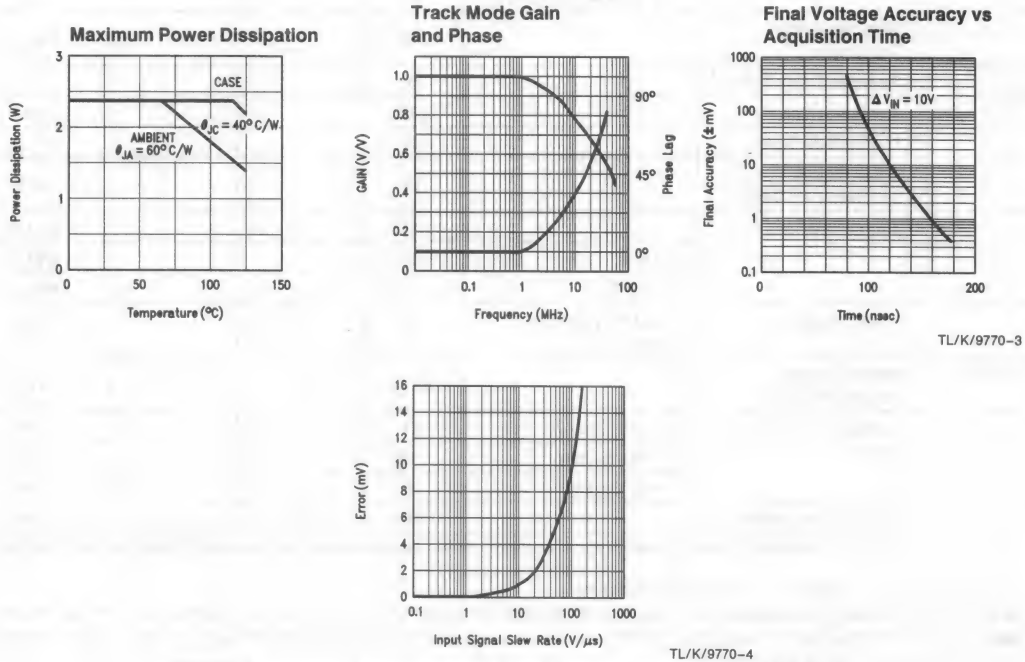


FIGURE 2. Accuracy Error Due to a ±50 ps Aperture Jitter at 10V Full Scale

Application Information

LAYOUT

The LH4860 is constructed in a way that with proper care in the layout it will meet its specifications without additional external components.

A large analog ground plane will provide uniform ground potential to the four ground pins (Pin 10, 15, 21 and 23). These pins should be connected to the ground plane with minimum lead length. Any difference in ground potential, due to ground current, will degrade the performance of the device.

The analog and digital grounds of the LH4860 should be connected together close to the device. The +5V digital logic supply needs to be well bypassed. Although both +5V and ±15V are internally decoupled with 0.001 μF, in critical applications, additional bypass capacitors are recommended (0.1 μF–1 μF tantalum).

LOGIC COMMANDS

A TTL logic "0" on Pin 11 (or a logic "1" on Pin 12) will put the LH4860 into the sample (track) mode. In this mode, the device acts as an inverting unity gain amplifier, and its output will track the input.

A logic "1" on Pin 11 and logic "0" on Pin 12 will put the device into the hold mode, where the output will be held constant at the level present when the command was given.

Application Information (Continued)

Unused logic pins need to be tied to a fixed logic level. When Pin 11 is used, then Pin 12 must be tied to ground; when Pin 12 is used as logic input, Pin 11 is to be tied to +5V through 1 kΩ.

Each Pin 11 or 12 represents one TTL load to the drive circuit.

Pin 11 (Hold)	Pin 12 (Hold)	State
0	0	Track
0	1	Track
1	0	Hold
1	1	Track

In the tracking mode, the Track-Hold Amplifier operates as an inverting amplifier with unity gain. It is limited by its small signal bandwidth, typically 16 MHz, and the power bandwidth, typically 4.8 MHz.

LOADING

Some restrictions on the output load apply to avoid oscillations and performance variations over temperature.

. . . Recommended load resistance is 500Ω or above and capacitance up to 50 pF; load resistance down to 250Ω can be used without degrading the performance. Capacitive loads up to 150 pF will be free of oscillations, but acquisition and settling times will be extended due to slew rate limitations in the output.

APERTURE JITTER

In a typical DSP Application, an analog signal needs to be digitized. This can be done with an A/D Converter; which has the limitation that the signal needs to be fairly constant throughout the conversion time, therefore, only low frequency signals can be converted without loss of accuracy. To handle faster signals, a Track-Hold Amplifier can be used in front of the A/D.

In order not to lose accuracy, the standard rule of thumb is that the input signal should not change more than ±½ LSB during the conversion time. This determines the maximum frequency for accurate conversion.

For example, take a 12-bit 10 μs A/D Converter. If it is operated on a 0V to 10V input range, 1 LSB is equivalent to:

$$\frac{10V}{2^{12}} = \frac{10V}{4096} = 2.44\ mV$$

and ½ LSB is 1.22 mV. The maximum allowable rate of change becomes:

$$\frac{dV}{dt} = \frac{½\ LSB}{Conversion\ Time} = \frac{1.22\ mV}{10\ \mu s} = 122\ V/s$$

For a sinewave of v(t) = A sine 2π ft, the derivative is a rate of change vs. time.

$$\frac{d\,v(t)}{dt} = 2\pi f\ A\ Cos\ (2\pi\ ft)$$

The extreme value of this is at t = 0, and the maximum rate of change becomes 2πfA.

If the sinewave is chosen for 10 V_{P-P}, or A = 5V, the maximum rate of change becomes 10πf. If this is equated to the maximum allowable rate of change, the frequency that can be converted accurately becomes:

$$f = \frac{122\ V/s}{2\pi 5V} = 3.9\ Hz$$

If a track-hold amplifier is used in front of the A/D, then much faster signals can be accurately digitized. In this case, the input waveform has to be repetitive, and the hold pulse is shifted in phase every time a new conversion is made, until the whole signal has been captured. The limitation for accuracy is determined by the aperture jitter, which is the uncertainty of the moment when the signal is frozen. In this case, the maximum slew rate is 1.22 mV/100 ps = 12.2 V/μs and the highest frequency at which accurate conversion occurs becomes:

$$f = \frac{12.2V/\mu s}{2\pi 5V} = 338\ kHz$$

The fact that the LH4860 can digitize the fastest part of a 338 kHz sine wave does not mean it can digitize that signal for reconstruction purposes. Realistically a sample can only be taken in the time it takes to acquire (200 ns for the LH4860) plus the conversion time of the ADC.

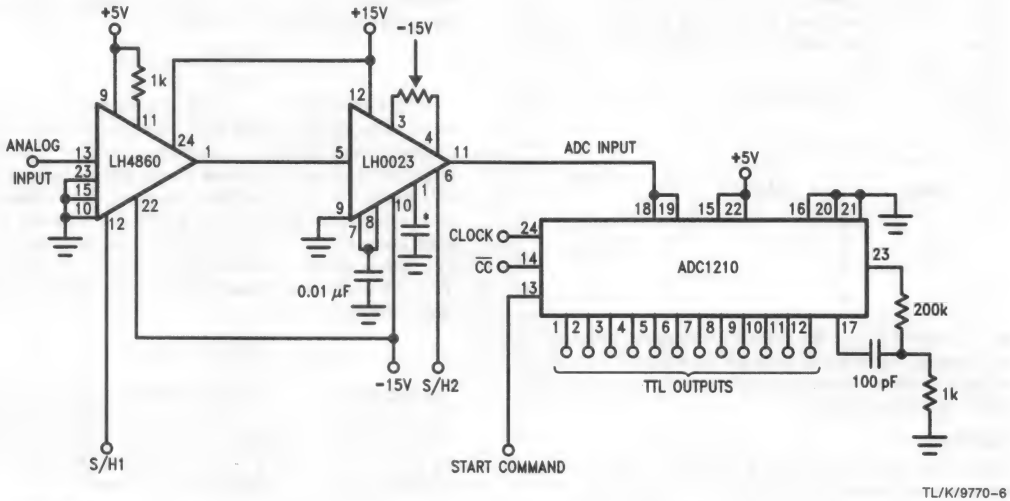
Other Considerations for Using the LH4860 with A/D Converters

There are several considerations for good match between track hold amplifier and A/D. One is that the output resistance of the T/H should be low compared to the input resistance of the A/D, up to frequencies 5 times the clockrate of the A/D. This is because of the digital nature of a successive approximation A/D its internal D/A changes its output momentarily and current transients occur at the A/D input. These should be sunk and settled before the next bit conversion. In the hold mode, the LH4860 has a typical output resistance of 0.1Ω; its output, typically, recovers to ±0.01% from 2 mA step in less than 100 ns.

Another consideration is the LH4860's track-to-hold transient settling time. Normally, the same timing pulse that initiates "hold" also starts the A/D conversion. The decision for the A/D's MSB, normally, takes place one clock cycle after the start signal, and at that time, the track-hold command pin can be driven directly (or inverted) from the successive approximation A/D's conversion status output. During conversion the T/H is in hold.

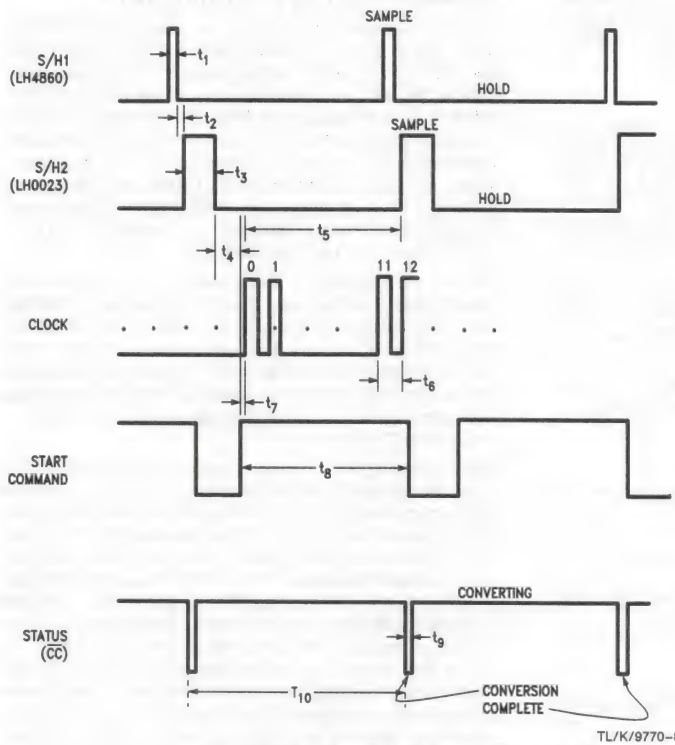
Many sampling A/D converter applications require that a signal be sampled fast but held for a long time so that a slow (inexpensive) A/D converter may be used. Such conflicting requirements place stringent demands on a S/H amplifier. Fortunately, cascading two S/H amplifiers, as in *Figure 3,* solves the problem. The LH4860 acquires the signal to within 0.01% F.S. in under 200 ns and holds it until the LH0023 acquires the sampled signal. The low droop rate of LH0023 allows it to hold the sampled signal to within 0.01% for as long as the conversion time of ADC1210 (100 μs). Note that the start pulse for the A/D converter should occur at the end of LH0023's hold mode settling time. *Figure 3's* circuit accepts a 0V to −5V full scale input signal and produces a complementary binary output. A typical timing dia-

Applications



*Polystyrene Cap 0.01 μF

FIGURE 3. Sampling A/D Converter



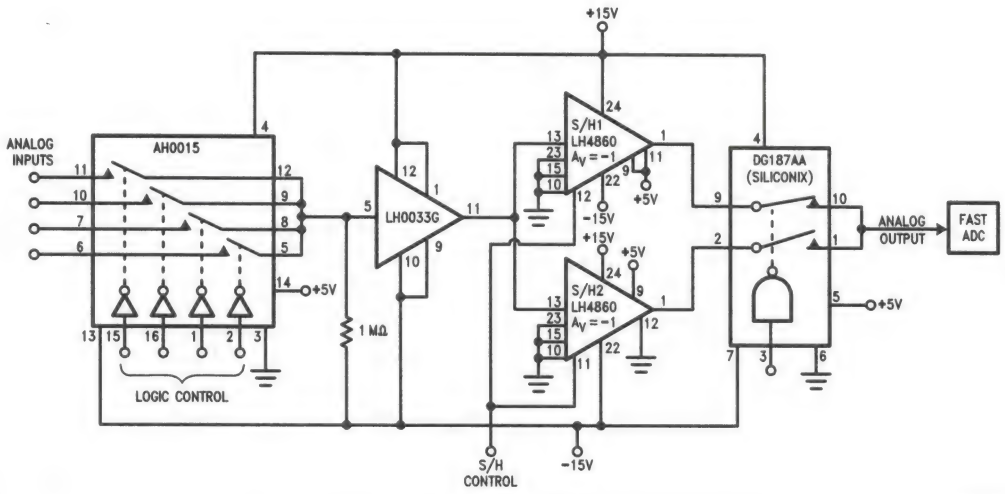
t_1 = S/H1 acquisition time = 200 ns
t_2 = S/H1 hold mode settling time = 100 ns
t_3 = S/H2 acquisition time = 20 μs
t_4 = S/H2 hold mode settling time = 15 μs
t_5 = Conversion time of ADC1210 = 100 μs
t_6 = Clock period = 7.7 μs
t_7 = Start pulse setup time = 30 ns
t_8 = Start pulse width = 105 μs
t_9 = Conversion complete time = 5 μs
t_{10} = Sampling period = $t_3 + t_4 + t_5 + t_7$ ≅ 135 μs

$$f_s = \frac{1}{t_{10}} \cong 7.4 \text{ kHz}$$

FIGURE 4. Typical Timing Diagram for the Sampling A/D Converter (Not Drawn to Scale)

gram for the sampling A/D converter *(Figure 4)* shows a sampling frequency of 7.4 kHz, thus from Nyquist criteria, the input signal's maximum frequency should be limited to 3.52 kHz (i.e., half the sampling frequency). This circuit is well suited for capturing fast single shot events as well as repetitive signals and for operation at elevated temperatures at which the increased droop rate of LH4860 is compensated for by the low droop rate of LH0023.

Applications (Continued)



TL/K/9770–7

FIGURE 5. Fast Data Acquisition Using Ping-Pong Switching

For ultrafast A/D converters with conversion time in the microsecond region, the S/H's acquisition and hold mode settling time can contribute significantly to the system's overall cycle time thus reducing system throughput. For example, an LH4860 at the front end of a 12-bit/1 μs ADC can add as much as 300 ns to the converter's conversion time thus increasing the sampling interval to 1300 ns; a 30% increase. However, by using two S/H amplifiers in a ping-pong switching configuration, the system's cycle time can be decreased to very nearly that of the ADC. *Figure 5* shows an ultrafast multi-channel data acquisition system. The AH0015 Mux allows the selection of 1 of 4 input signals in a process monitor application. Note that a logic "1" at AH0015's logic control pin closes the corresponding switch. Since LH4860's input impedance is only 1 kΩ, LH0033 buffers the input signal so as to prevent the S/H from loading the signal source, a 1 MΩ resistor at the buffer's input prevents the buffer's output from saturating when all Mux switches are open. The ping-pong switching scheme involves the use of a fast SPDT switch (DG187AA) to select the output of each S/H for half the sampling period. Thus, S/H1's output is selected when S/H1 is in the hold mode, the ADC meanwhile begins conversion. While the ADC is converting, S/H2 is acquiring a new sample of the input signal. As soon as the ADC's conversion is complete, S/H2 goes into hold mode and a new conversion can begin at the end of S/H2's hold mode settling time, S/H1 now goes into the sampling mode. This approach thus eliminates the S/H's acquisition time from the system's overall cycle time. Note that DG187AA's typical switch on and off times are less than the 100 ns hold mode settling time of the LH4860. Consequently, switching the S/H's output to the output channel at the instant the S/H goes from sample to hold mode eliminates the switch delay because the ADC's conversion does not begin until after the LH4860's hold mode settling time of 100 ns. Neglecting the minimal delay through LH0033G, a sampling interval of 1100 ns can be achieved, this is only a 10% increase over the minimum available sampling interval. It should be noted that since LH0033 and LH4860 do not have precisely unity gain, they introduce gain error in addition to voltage offset. The gain error and offset of the entire system can, however, be trimmed out by the ADC's gain and offset trim circuits. The circuit in *Figure 5* accepts −5V to +5V input signals and produces an inverted output. The circuit described is ideally suited for interface with flash ADCs having sub-microsecond conversion times, and, taking advantage of the ping-pong switching scheme allows significant improvement in system throughput compared to a single S/H and ADC combination.

Section 6

Temperature Sensors

Section 6 Contents

Temperature Sensors Selection Guide 6-3
LM34/LM34A/LM34C/LM34CA/LM34D Precision Fahrenheit Temperature Sensors 6-4
LM35/LM35A/LM35C/LM35CA/LM35D Precision Centigrade Temperature Sensors 6-12
LM135/LM235/LM335/LM135A/LM235A/LM335A Precision Temperature Sensors 6-21
LM3911 Temperature Controller 6-30

National Semiconductor

Temperature Sensor Selection Guide

Part	Temp. Range	*Accuracy	Output Scale
LM34A	−50°F to +300°F	±2.0°F	10 mV/°F
LM34	−50°F to +300°F	±3.0°F	10 mV/°F
LM34CA	−40°F to +230°F	±2.0°F	10 mV/°F
LM34C	−40°F to +230°F	±3.0°F	10 mV/°F
LM34D	+32°F to +212°F	±4.0°F	10 mV/°F
LM35A	−55°C to +150°C	±1.0°C	10 mV/°C
LM35	−55°C to +150°C	±1.5°C	10 mV/°C
LM35CA	−40°C to +110°C	±1.0°C	10 mV/°C
LM35C	−40°C to +110°C	±1.5°C	10 mV/°C
LM35D	0°C to +100°C	±2.0°C	10 mV/°C
LM134-3	−55°C to +125°C	±3.0°C	$I_{SET} \propto$ °k
LM134-6	−55°C to +125°C	±6.0°C	$I_{SET} \propto$ °k
LM234-3	−25°C to +100°C	±3.0°C	$I_{SET} \propto$ °k
LM234-6	−25°C to +100°C	±6.0°C	$I_{SET} \propto$ °k
LM135A	−55°C to +150°C	±1.3°C	10 mV/°k
LM135	−55°C to +150°C	±2.0°C	10 mV/°k
LM235A	−40°C to +125°C	±1.3°C	10 mV/°k
LM235	−40°C to +125°C	±2.0°C	10 mV/°k
LM335A	−40°C to +100°C	±2.0°C	10 mV/°k
LM335	−40°C to +100°C	±4.0°C	10 mV/°k
LM3911	−25°C to +85°C	±10.0°C	10 mV/°k (or °F)

***Note:** Accuracy is measured over T(Min) to T(Max) uncalibrated

Note: The LM134/234/334 3-Terminal Adjustable current sources Datasheet can be found in Linear 1, Section 1.

National Semiconductor

LM34/LM34A/LM34C/LM34CA/LM34D Precision Fahrenheit Temperature Sensors

General Description

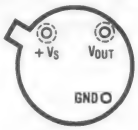
The LM34 series are precision integrated-circuit temperature sensors, whose output voltage is linearly proportional to the Fahrenheit temperature. The LM34 thus has an advantage over linear temperature sensors calibrated in degrees Kelvin, as the user is not required to subtract a large constant voltage from its output to obtain convenient Fahrenheit scaling. The LM34 does not require any external calibration or trimming to provide typical accuracies of ±½°F at room temperature and ±1½°F over a full −50 to +300°F temperature range. Low cost is assured by trimming and calibration at the wafer level. The LM34's low output impedance, linear output, and precise inherent calibration make interfacing to readout or control circuitry especially easy. It can be used with single power supplies or with plus and minus supplies. As it draws only 70 μA from its supply, it has very low self-heating, less than 0.2°F in still air. The LM34 is rated to operate over a −50° to +300°F temperature range, while the LM34C is rated for a −40° to +230°F range (0°F with improved accuracy). The LM34 series is available packaged in hermetic TO-46 transistor packages, while the LM34C is also available in the plastic TO-92 transistor package. The LM34 is a complement to the LM35 (Centigrade) temperature sensor.

Features

- Calibrated directly in degrees Fahrenheit
- Linear +10.0 mV/°F scale factor
- 1.0°F accuracy guaranteed (at +77°F)
- Rated for full −50° to +300°F range
- Suitable for remote applications
- Low cost due to wafer-level trimming
- Operates from 5 to 30 volts
- Less than 70 μA current drain
- Low self-heating, 0.18°F in still air
- Nonlinearity only ±0.5°F typical
- Low-impedance output, 0.4Ω for 1 mA load

Connection Diagrams

TO-46
Metal Can Package*

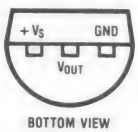


TL/H/6685–1

*Case is connected to negative pin.

Order Numbers LM34H, LM34AH, LM34CH, LM34CAH or LM34DH
See NS Package Number H03H

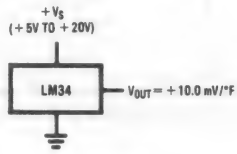
TO-92
Plastic Package



TL/H/6685–2

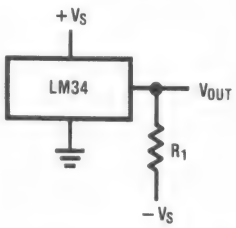
Order Number LM34CZ or LM34DZ
See NS Package Number Z03A

Typical Applications



TL/H/6685–3

FIGURE 1. Basic Fahrenheit Temperature Sensor (+5° to +300°F)



CHOOSE $R_1 = (-V_S)/50\ \mu A$
$V_{OUT} = +3{,}000$ mV AT +300°F
$= +750$ mV AT +75°F
$= -500$ mV AT −50°F

TL/H/6685–4

FIGURE 2. Full-Range Fahrenheit Temperature Sensor

Absolute Maximum Ratings (Note 10)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	+35V to −0.2V
Output Voltage	+6V to −1.0V
Output Current	10 mA
Storage Temperature,	
TO-46 Package	−76°F to +356°F
TO-92 Package	−76°F to +300°F
Lead Temp. (Soldering, 10 seconds)	
TO-46 Package	+300°C
TO-92 Package	+260°C
Specified Operating Temp. Range (Note 2)	
	T_{MIN} to T_{MAX}
LM34, LM34A	−50°F to +300°F
LM34C, LM34CA	−40°F to +230°F
LM34D	+32°F to +212°F

DC Electrical Characteristics (Note 1, **Note 6**)

Parameter	Conditions	LM34A			LM34CA			Units (Max)
		Typical	Tested Limit (Note 4)	Design Limit (Note 5)	Typical	Tested Limit (Note 4)	Design Limit (Note 5)	
Accuracy (Note 7)	T_A = +77°F	±0.4	±1.0		±0.4	±1.0		°F
	T_A = 0°F	±0.6			±0.6		±2.0	°F
	T_A = T_{MAX}	±0.8	±2.0		±0.8	±2.0		°F
	T_A = T_{MIN}	±0.8	±2.0		±0.8		±3.0	°F
Nonlinearity (Note 8)	$T_{MIN} \le T_A \le T_{MAX}$	**±0.35**		**±0.7**	**±0.30**		**±0.6**	°F
Sensor Gain (Average Slope)	$T_{MIN} \le T_A \le T_{MAX}$	**+10.0**	**+9.9, +10.1**		**+10.0**		**+9.9, +10.1**	mV/°F, min mV/°F, max
Load Regulation (Note 3)	T_A = +77°F $T_{MIN} \le T_A \le T_{MAX}$ $0 \le I_L \le 1$ mA	±0.4 **±0.5**	±1.0	 **±3.0**	±0.4 **±0.5**	±1.0	 **±3.0**	mV/mA mV/mA
Line Regulation (Note 3)	T_A = +77°F 5V ≤ V_S ≤ 30V	±0.01 **±0.02**	±0.05	 **±0.1**	±0.01 **±0.02**	±0.05	 **±0.1**	mV/V mV/V
Quiescent Current (Note 9)	V_S = +5V, +77°F V_S = +5V V_S = +30V, +77°F V_S = +30V	75 **131** 76 **132**	90 92 	 **160** **163**	75 **116** 76 **117**	90 92 	 **139** **142**	µA µA µA µA
Change of Quiescent Current (Note 3)	4V ≤ V_S ≤ 30V, +77°F 5V ≤ V_S ≤ 30V	+0.5 **+1.0**	2.0	 **3.0**	0.5 **1.0**	2.0	 **3.0**	µA µA
Temperature Coefficient of Quiescent Current		**+0.30**		**+0.5**	**+0.30**		**+0.5**	µA/°F
Minimum Temperature for Rated Accuracy	In circuit of *Figure 1*, I_L = 0	+3.0		+5.0	+3.0		+5.0	°F
Long-Term Stability	T_j = T_{MAX} for 1000 hours	±0.16			±0.16			°F

Note 1: Unless otherwise noted, these specifications apply: −50°F ≤ T_j ≤ + 300°F for the LM34 and LM34A; −40°F ≤ T_j ≤ +230°F for the LM34C and LM34CA; and +32°F ≤ T_j ≤ + 212°F for the LM34D. V_S = +5 Vdc and I_{LOAD} = 50 µA in the circuit of *Figure 2;* +6 Vdc for LM34 and LM34A for 230°F ≤ T_j ≤ 300°F. These specifications also apply from +5°F to T_{MAX} in the circuit of *Figure 1.*

Note 2: Thermal resistance of the TO-46 package is 92°F/W junction to ambient and 43°F/W junction to case. Thermal resistance of the TO-92 package is 324°F/W junction to ambient.

Note 3: Regulation is measured at constant junction temperature using pulse testing with a low duty cycle. Changes in output due to heating effects can be computed by multiplying the internal dissipation by the thermal resistance.

Note 4: Tested limits are guaranteed and 100% tested in production.

Note 5: Design limits are guaranteed (but not 100% production tested) over the indicated temperature and supply voltage ranges. These limits are not used to calculate outgoing quality levels.

Note 6: Specification in **BOLDFACE TYPE** apply over the full rated temperature range.

Note 7: Accuracy is defined as the error between the output voltage and 10 mV/°F times the device's case temperature at specified conditions of voltage, current, and temperature (expressed in °F).

Note 8: Nonlinearity is defined as the deviation of the output-voltage-versus-temperature curve from the best-fit straight line over the device's rated temperature range.

Note 9: Quiescent current is defined in the circuit of *Figure 1.*

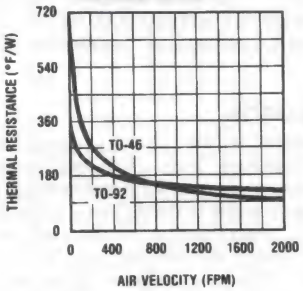
Note 10: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its rated operating conditions (see Note 1).

DC Electrical Characteristics (Note 1, **Note 6**) (Continued)

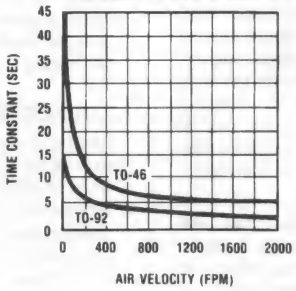
Parameter	Conditions	LM34			LM34C, LM34D			Units (Max)
		Typical	Tested Limit (Note 4)	Design Limit (Note 5)	Typical	Tested Limit (Note 4)	Design Limit (Note 5)	
Accuracy, LM34, LM34C (Note 7)	$T_A = +77°F$	±0.8	±2.0		±0.8	±2.0		°F
	$T_A = 0°F$	±1.0			±1.0		±3.0	°F
	$T_A = T_{MAX}$	±1.6	±3.0		±1.6		±3.0	°F
	$T_A = T_{MIN}$	±1.6		±3.0	±1.6		±4.0	°F
Accuracy, LM34D (Note 7)	$T_A = +77°F$				±1.2	±3.0		°F
	$T_A = T_{MAX}$				±1.8		±4.0	°F
	$T_A = T_{MIN}$				±1.8		±4.0	°F
Nonlinearity (Note 8)	$T_{MIN} \le T_A \le T_{MAX}$	**±0.6**		**±1.0**	**±0.4**		**±1.0**	°F
Sensor Gain (Average Slope)	$T_{MIN} \le T_A \le T_{MAX}$	**+10.0**	**+9.8, +10.2**		**+10.0**		**+9.8, +10.2**	mV/°F, min mV/°F, max
Load Regulation (Note 3)	$T_A = +77°F$	±0.4	±2.5		±0.4	±2.5		mV/mA
	$T_{MIN} \le T_A \le +150°F$	**±0.5**		**±6.0**	**±0.5**		**±6.0**	mV/mA
	$0 \le I_L \le 1$ mA							
Line Regulation (Note 3)	$T_A = +77°F$	±0.01	±0.1		±0.01	±0.1		mV/V
	$5V \le V_S \le 30V$	**±0.02**		**±0.2**	**±0.02**		**±0.2**	mV/V
Quiescent Current (Note 9)	$V_S = +5V, +77°F$	75	100		75	100		µA
	$V_S = +5V$	**131**		**176**	**116**		**154**	µA
	$V_S = +30V, +77°F$	76	103		76	103		µA
	$V_S = +30V$	**132**		**181**	**117**		**159**	µA
Change of Quiescent Current (Note 3)	$4V \le V_S \le 30V, +77°F$	+0.5	3.0		0.5	3.0		µA
	$5V \le V_S \le 30V$	**+1.0**		**5.0**	**1.0**		**5.0**	µA
Temperature Coefficient of Quiescent Current		**+0.30**		**+0.7**	**+0.30**		**+0.7**	µA/°F
Minimum Temperature for Rated Accuracy	In circuit of *Figure 1*, $I_L = 0$	+3.0		+5.0	+3.0		+5.0	°F
Long-Term Stability	$T_j = T_{MAX}$ for 1000 hours	±0.16			±0.16			°F

Typical Performance Characteristics

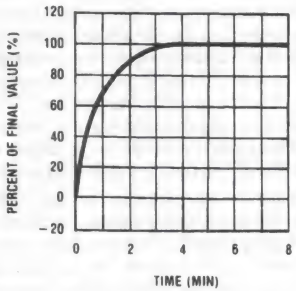
Thermal Resistance Junction to Air



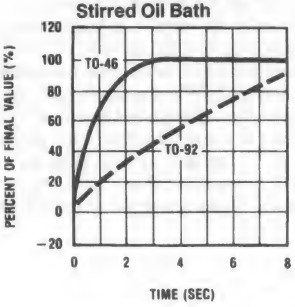
Thermal Time Constant



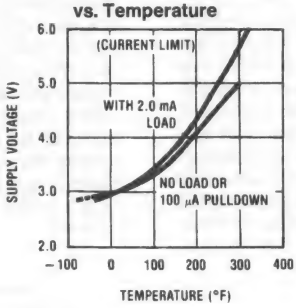
Thermal Response in Still Air



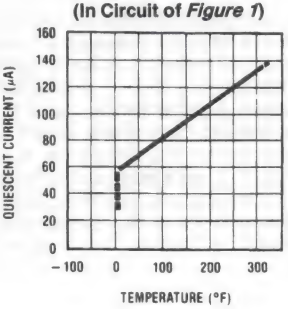
Thermal Response in Stirred Oil Bath



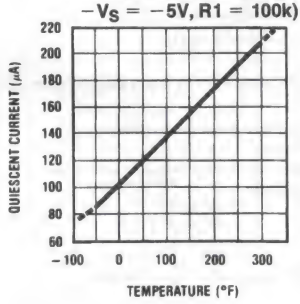
Minimum Supply Voltage vs. Temperature



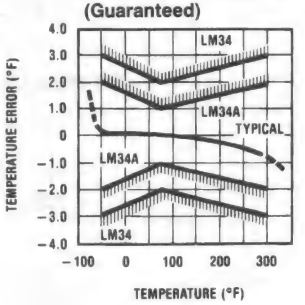
Quiescent Current vs. Temperature (In Circuit of *Figure 1*)



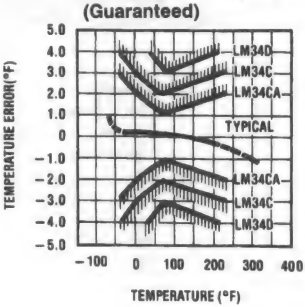
Quiescent Current vs. Temperature (In Circuit of *Figure 2*; $-V_S = -5V$, R1 = 100k)



Accuracy vs. Temperature (Guaranteed)



Accuracy vs. Temperature (Guaranteed)



TL/H/6685–5

Typical Applications

The LM34 can be applied easily in the same way as other integrated-circuit temperature sensors. It can be glued or cemented to a surface and its temperature will be within about 0.02°F of the surface temperature. This presumes that the ambient air temperature is almost the same as the surface temperature; if the air temperature were much higher or lower than the surface temperature, the actual temperature of the LM34 die would be at an intermediate temperature between the surface temperature and the air temperature. This is expecially true for the TO-92 plastic package, where the copper leads are the principal thermal path to carry heat into the device, so its temperature might be closer to the air temperature than to the surface temperature.

To minimize this problem, be sure that the wiring to the LM34, as it leaves the device, is held at the same temperature as the surface of interest. The easiest way to do this is to cover up these wires with a bead of epoxy which will insure that the leads and wires are all at the same temperature as the surface, and that the LM34 die's temperature will not be affected by the air temperature.

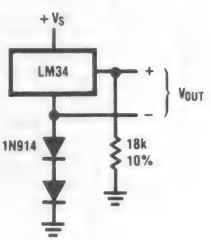
The TO-46 metal package can also be soldered to a metal surface or pipe without damage. Of course in that case, the V_- terminal of the circuit will be grounded to that metal. Alternatively, the LM34 can be mounted inside a sealed-end metal tube, and can then be dipped into a bath or screwed into a threaded hole in a tank. As with any IC, the LM34 and accompanying wiring and circuits must be kept insulated and dry, to avoid leakage and corrosion. This is especially true if the circuit may operate at cold temperatures where condensation can occur. Printed-circuit coatings and varnishes such as Humiseal and epoxy paints or dips are often used to insure that moisture cannot corrode the LM34 or its connections.

These devices are sometimes soldered to a small, lightweight heat fin to decrease the thermal time constant and speed up the response in slowly-moving air. On the other hand, a small thermal mass may be added to the sensor to give the steadiest reading despite small deviations in the air temperature.

Capacitive Loads

Like most micropower circuits, the LM34 has a limited ability to drive heavy capacitive loads. The LM34 by itself is able to drive 50 pF without special precautions. If heavier loads are anticipated, it is easy to isolate or decouple the load with a resistor; see *Figure 3*. Or you can improve the tolerance of capacitance with a series R-C damper from output to ground; see *Figure 4*. When the LM34 is applied with a 499Ω load resistor (as shown), it is relatively immune to wiring capacitance because the capacitance forms a bypass from ground to input, not on the output. However, as with any linear circuit connected to wires in a hostile environment, its performance can be affected adversely by intense electromagnetic sources such as relays, radio transmitters, motors with arcing brushes, SCR's transients, etc., as its wiring can act as a receiving antenna and its internal junctions can act as rectifiers. For best results in such cases, a bypass capacitor from V_{IN} to ground and a series R-C damper such as 75Ω in series with 0.2 or 1 μF from output to ground are often useful. These are shown in the following circuits.

Temperature Sensor, Single Supply, −50° to +300°F



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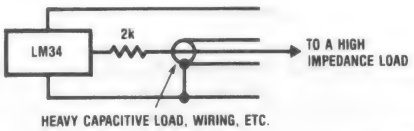

TL/H/6685–7

FIGURE 3. LM34 with Decoupling from Capacitive Load

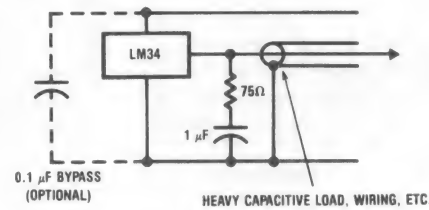

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FIGURE 4. LM34 with R-C Damper

Temperature Rise of LM34 Due to Self-Heating (Thermal Resistance)

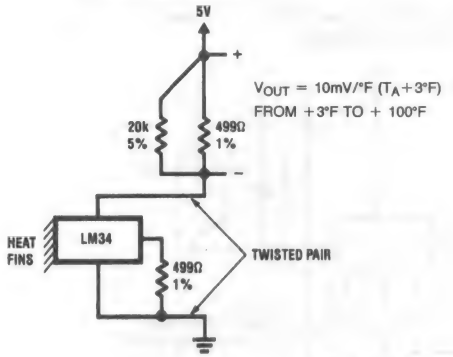
Conditions	TO-46, No Heat Sink	TO-46, Small Heat Fin*	TO-92, No Heat Sink	TO-92, Small Heat Fin**
Still air	720°F/W	180°F/W	324°F/W	252°F/W
Moving air	180°F/W	72°F/W	162°F/W	126°F/W
Still oil	180°F/W	72°F/W	162°F/W	126°F/W
Stirred oil	90°F/W	54°F/W	81°F/W	72°F/W
(Clamped to metal, infinite heat sink)	(43°F/W)			

*Wakefield type 201 or 1" disc of 0.020" sheet brass, soldered to case, or similar.

**TO-92 package glued and leads soldered to 1" square of 1/16" printed circuit board with 2 oz copper foil, or similar.

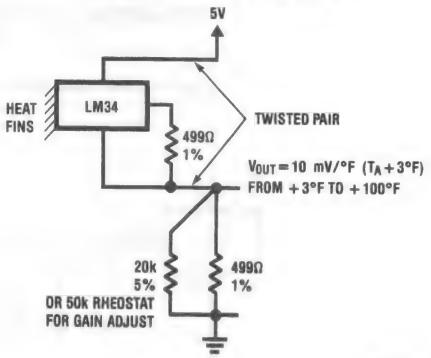
Typical Applications (Continued)

Two-Wire Remote Temperature Sensor (Grounded Sensor)



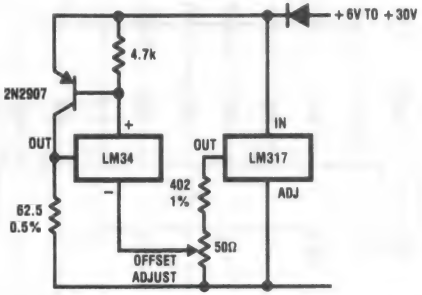
TL/H/6685–9

Two-Wire Remote Temperature Sensor (Output Referred to Ground)



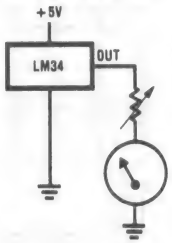
TL/H/6685–10

4-to-20 mA Current Source (0 to + 100°F)



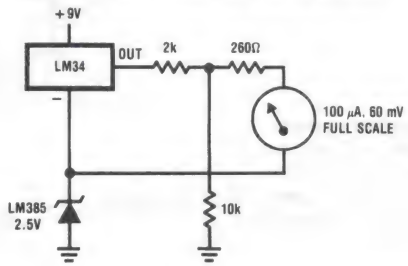
TL/H/6685–11

Fahrenheit Thermometer (Analog Meter)



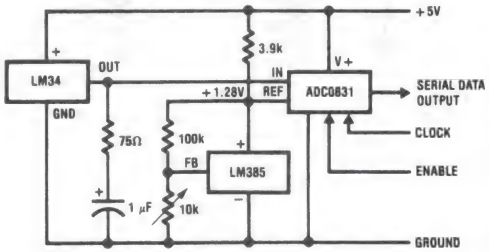
TL/H/6685–12

Expanded Scale Thermometer (50° to 80° Fahrenheit, for Example Shown)



TL/H/6685–13

Temperature-to-Digital Converter (Serial Output, + 128°F Full Scale)



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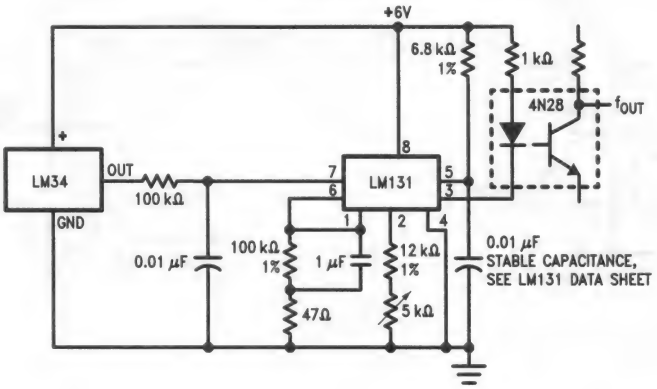
6

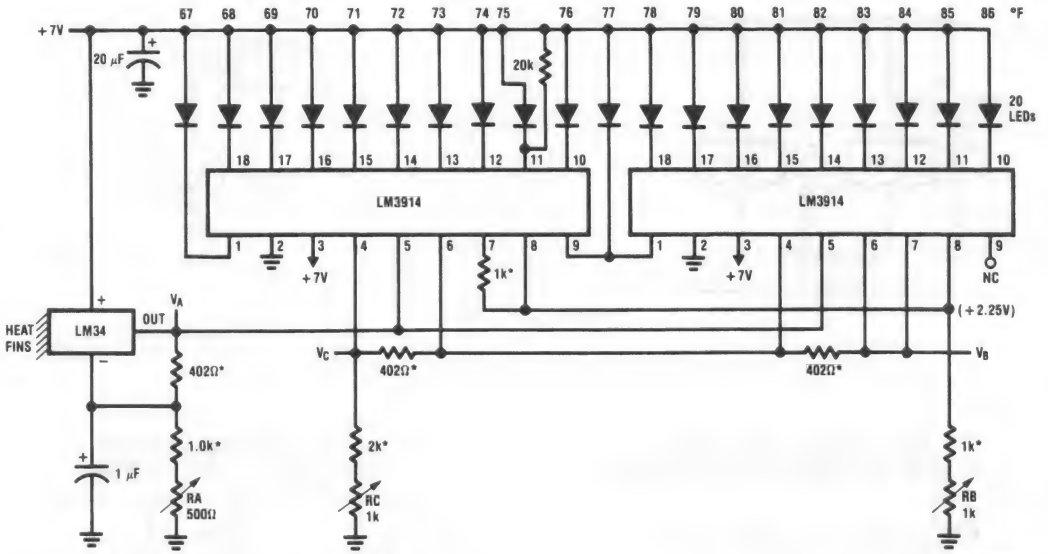
Typical Applications (Continued)

LM34 with Voltage-to-Frequency Converter and Isolated Output (3°F to + 300°F; 30 Hz to 3000 Hz)



TL/H/6685–15

Bar-Graph Temperature Display (Dot Mode)

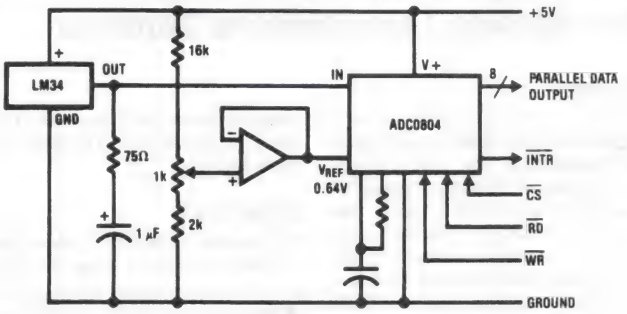


TL/H/6685–16

* = 1% or 2% film resistor
—Trim R_B for V_B = 3.525V
—Trim R_C for V_C = 2.725V
—Trim R_A for V_A = 0.085V + 40 mV/°F × $T_{AMBIENT}$
—Example, V_A = 3.285V at 80°F

Typical Applications (Continued)

Temperature-to-Digital Converter
(Parallel TRI-STATE® Outputs for Standard Data Bus to μP Interface, 128 °F Full Scale)



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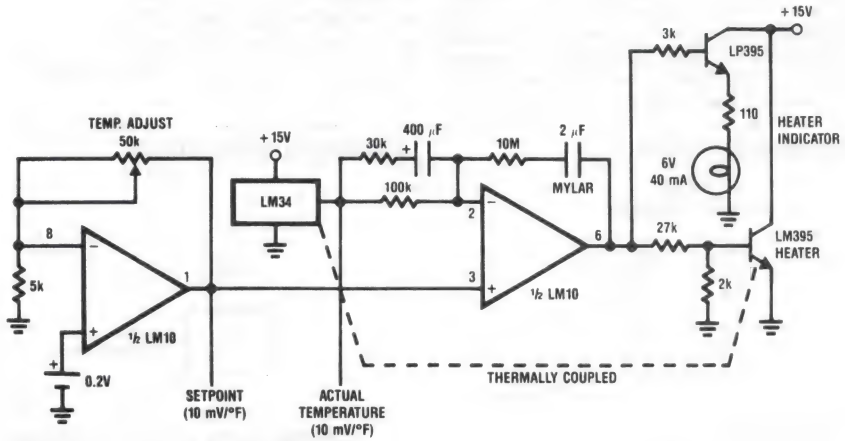
Temperature Controller



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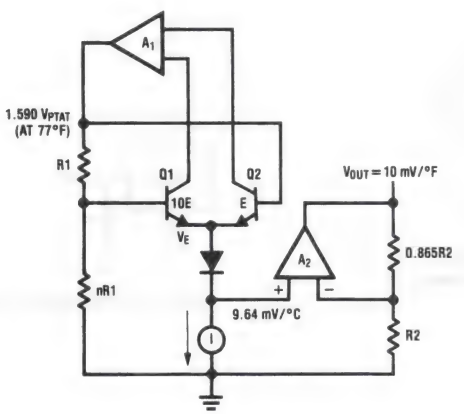
Block Diagram



TL/H/6685–19

National Semiconductor

LM35/LM35A/LM35C/LM35CA/LM35D Precision Centigrade Temperature Sensors

General Description

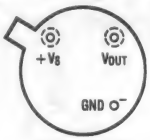
The LM35 series are precision integrated-circuit temperature sensors, whose output voltage is linearly proportional to the Celsius (Centigrade) temperature. The LM35 thus has an advantage over linear temperature sensors calibrated in ° Kelvin, as the user is not required to subtract a large constant voltage from its output to obtain convenient Centigrade scaling. The LM35 does not require any external calibration or trimming to provide typical accuracies of ±¼°C at room temperature and ±¾°C over a full −55 to +150°C temperature range. Low cost is assured by trimming and calibration at the wafer level. The LM35's low output impedance, linear output, and precise inherent calibration make interfacing to readout or control circuitry especially easy. It can be used with single power supplies, or with plus and minus supplies. As it draws only 60 μA from its supply, it has very low self-heating, less than 0.1°C in still air. The LM35 is rated to operate over a −55° to +150°C temperature range, while the LM35C is rated for a −40° to +110°C range (−10° with improved accuracy). The LM35 series is available packaged in hermetic TO-46 transistor packages, while the LM35C is also available in the plastic TO-92 transistor package.

Features

- Calibrated directly in ° Celsius (Centigrade)
- Linear + 10.0 mV/°C scale factor
- 0.5°C accuracy guaranteeable (at +25°C)
- Rated for full −55° to +150°C range
- Suitable for remote applications
- Low cost due to wafer-level trimming
- Operates from 4 to 30 volts
- Less than 60 μA current drain
- Low self-heating, 0.08°C in still air
- Nonlinearity only ±¼°C typical
- Low impedance output, 0.1 Ω for 1 mA load

Connection Diagrams

TO-46
Metal Can Package*



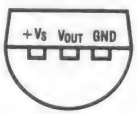
BOTTOM VIEW

TL/H/5516–1

*Case is connected to negative pin

Order Number LM35H, LM35AH, LM35CH, LM35CAH or LM35DH
See NS Package Number H03H

TO-92
Plastic Package



BOTTOM VIEW

TL/H/5516–2

Order Number LM35CZ or LM35DZ
See NS Package Number Z03A

Typical Applications

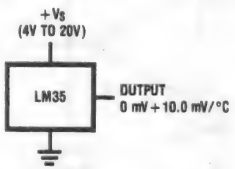


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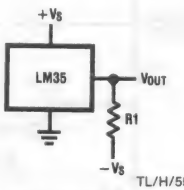
FIGURE 1. Basic Centigrade Temperature Sensor (+2°C to +150°C)



Choose $R_1 = -V_S/50\ \mu A$

$V_{OUT} = +1{,}500$ mV at +150°C
$= +250$ mV at +25°C
$= -550$ mV at −55°C

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FIGURE 2. Full-Range Centigrade Temperature Sensor

Absolute Maximum Ratings (Note 10)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	+35V to −0.2V
Output Voltage	+6V to −1.0V
Output Current	10 mA
Storage Temp., TO-46 Package,	−60°C to +180°C
TO-92 Package,	−60°C to +150°C
Lead Temp. (Soldering, 10 seconds):	
TO-46 Package,	300°C
TO-92 Package,	260°C

Specified Operating Temperature Range: T_{MIN} to T_{MAX} (Note 2)

LM35, LM35A	−55°C to +150°C
LM35C, LM35CA	−40°C to +110°C
LM35D	0°C to +100°C

Electrical Characteristics (Note 1) (Note 6)

Parameter	Conditions	LM35A			LM35CA			Units (Max.)
		Typical	Tested Limit (Note 4)	Design Limit (Note 5)	Typical	Tested Limit (Note 4)	Design Limit (Note 5)	
Accuracy (Note 7)	$T_A=+25°C$	±0.2	±0.5		±0.2	±0.5		°C
	$T_A=-10°C$	±0.3			±0.3		±1.0	°C
	$T_A=T_{MAX}$	±0.4	±1.0		±0.4	±1.0		°C
	$T_A=T_{MIN}$	±0.4	±1.0		±0.4		±1.5	°C
Nonlinearity (Note 8)	$T_{MIN}\le T_A\le T_{MAX}$	**±0.18**		**±0.35**	**±0.15**		**±0.3**	°C
Sensor Gain (Average Slope)	$T_{MIN}\le T_A\le T_{MAX}$	**+10.0**	**+9.9, +10.1**		**+10.0**		**+9.9, +10.1**	mV/°C
Load Regulation (Note 3) $0\le I_L\le 1$ mA	$T_A=+25°C$	±0.4	±1.0		±0.4	±1.0		mV/mA
	$T_{MIN}\le T_A\le T_{MAX}$	**±0.5**		**±3.0**	**±0.5**		**±3.0**	mV/mA
Line Regulation (Note 3)	$T_A=+25°C$	±0.01	±0.05		±0.01	±0.05		mV/V
	$4V\le V_S\le 30V$	**±0.02**		**±0.1**	**±0.02**		**±0.1**	mV/V
Quiescent Current (Note 9)	$V_S=+5V$, +25°C	56	67		56	67		µA
	$V_S=+5V$	**105**		**131**	**91**		**114**	µA
	$V_S=+30V$, +25°C	56.2	68		56.2	68		µA
	$V_S=+30V$	**105.5**		**133**	**91.5**		**116**	µA
Change of Quiescent Current (Note 3)	$4V\le V_S\le 30V$, +25°C	0.2	1.0		0.2	1.0		µA
	$4V\le V_S\le 30V$	**0.5**		**2.0**	**0.5**		**2.0**	µA
Temperature Coefficient of Quiescent Current		**+0.39**		**+0.5**	**+0.39**		**+0.5**	µA/°C
Minimum Temperature for Rated Accuracy	In circuit of *Figure 1*, $I_L=0$	+1.5		+2.0	+1.5		+2.0	°C
Long Term Stability	$T_J=T_{MAX}$, for 1000 hours	±0.08			±0.08			°C

Note 1: Unless otherwise noted, these specifications apply: $-55°C\le T_J\le +150°C$ for the LM35 and LM35A; $-40°\le T_J\le +110°C$ for the LM35C and LM35CA; and $0°\le T_J\le +100°C$ for the LM35D. $V_S=+5$Vdc and $I_{LOAD}=50$ µA, in the circuit of *Figure 2*. These specifications also apply from +2°C to T_{MAX} in the circuit of *Figure 1*. Specifications in **boldface** apply over the full rated temperature range.

Note 2: Thermal resistance of the TO-46 package is 440°C/W, junction to ambient, and 24°C/W junction to case. Thermal resistance of the TO-92 package is 180°C/W junction to ambient.

Electrical Characteristics (Note 1) **(Note 6)** (Continued)

Parameter	Conditions	LM35			LM35C, LM35D			Units (Max.)
		Typical	Tested Limit (Note 4)	Design Limit (Note 5)	Typical	Tested Limit (Note 4)	Design Limit (Note 5)	
Accuracy, LM35, LM35C (Note 7)	$T_A = +25°C$	±0.4	±1.0		±0.4	±1.0		°C
	$T_A = -10°C$	±0.5			±0.5		±1.5	°C
	$T_A = T_{MAX}$	±0.8	±1.5		±0.8		±1.5	°C
	$T_A = T_{MIN}$	±0.8		±1.5	±0.8		±2.0	°C
Accuracy, LM35D (Note 7)	$T_A = +25°C$				±0.6	±1.5		°C
	$T_A = T_{MAX}$				±0.9		±2.0	°C
	$T_A = T_{MIN}$				±0.9		±2.0	°C
Nonlinearity (Note 8)	$T_{MIN} \le T_A \le T_{MAX}$	**±0.3**		**±0.5**	**±0.2**		**±0.5**	°C
Sensor Gain (Average Slope)	$T_{MIN} \le T_A \le T_{MAX}$	**+10.0**	**+9.8, +10.2**		**+10.0**		**+9.8, +10.2**	mV/°C
Load Regulation (Note 3) $0 \le I_L \le 1$ mA	$T_A = +25°C$	±0.4	±2.0		±0.4	±2.0		mV/mA
	$T_{MIN} \le T_A \le T_{MAX}$	**±0.5**		**±5.0**	**±0.5**		**±5.0**	mV/mA
Line Regulation (Note 3)	$T_A = +25°C$	±0.01	±0.1		±0.01	±0.1		mV/V
	$4V \le V_S \le 30V$	**±0.02**		**±0.2**	**±0.02**		**±0.2**	mV/V
Quiescent Current (Note 9)	$V_S = +5V, +25°C$	56	80		56	80		µA
	$V_S = +5V$	**105**		**158**	**91**		**138**	µA
	$V_S = +30V, +25°C$	56.2	82		56.2	82		µA
	$V_S = +30V$	**105.5**		**161**	**91.5**		**141**	µA
Change of Quiescent Current (Note 3)	$4V \le V_S \le 30V, +25°C$	0.2	2.0		0.2	2.0		µA
	$4V \le V_S \le 30V$	**0.5**		**3.0**	**0.5**		**3.0**	µA
Temperature Coefficient of Quiescent Current		**+0.39**		**+0.7**	**+0.39**		**+0.7**	µA/°C
Minimum Temperature for Rated Accuracy	In circuit of *Figure 1*, $I_L = 0$	+1.5		+2.0	+1.5		+2.0	°C
Long Term Stability	$T_J = T_{MAX}$, for 1000 hours	±0.08			±0.08			°C

Note 3: Regulation is measured at constant junction temperature, using pulse testing with a low duty cycle. Changes in output due to heating effects can be computed by multiplying the internal dissipation by the thermal resistance.

Note 4: Tested Limits are guaranteed and 100% tested in production.

Note 5: Design Limits are guaranteed (but not 100% production tested) over the indicated temperature and supply voltage ranges. These limits are not used to calculate outgoing quality levels.

Note 6: Specifications in **boldface** apply over the full rated temperature range.

Note 7: Accuracy is defined as the error between the output voltage and 10mv/°C times the device's case temperature, at specified conditions of voltage, current, and temperature (expressed in °C).

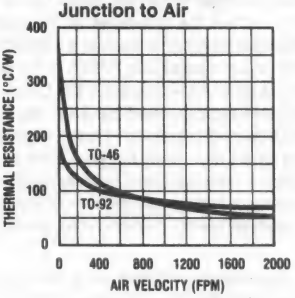
Note 8: Nonlinearity is defined as the deviation of the output-voltage-versus-temperature curve from the best-fit straight line, over the device's rated temperature range.

Note 9: Quiescent current is defined in the circuit of *Figure 1*.

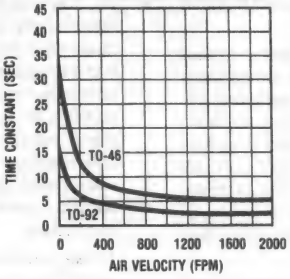
Note 10: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its rated operating conditions. See Note 1.

Typical Performance Characteristics

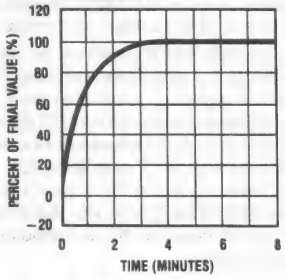
Thermal Resistance Junction to Air



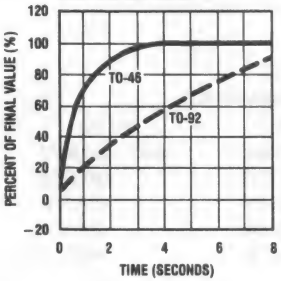
Thermal Time Constant



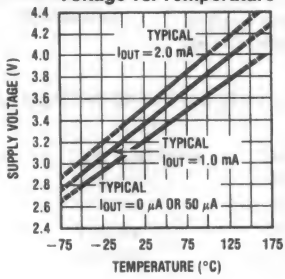
Thermal Response in Still Air



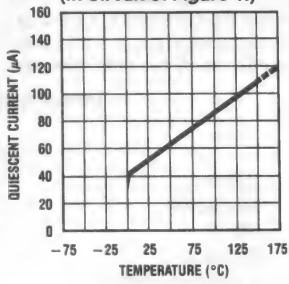
Thermal Response in Stirred Oil Bath



Minimum Supply Voltage vs. Temperature



Quiescent Current vs. Temperature (In Circuit of *Figure 1*.)



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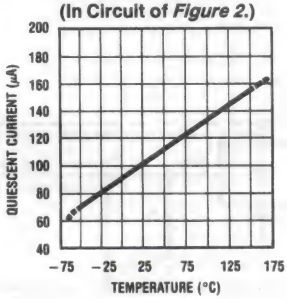
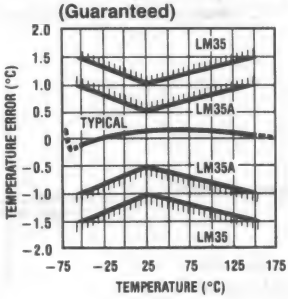
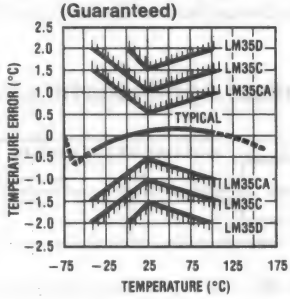
Quiescent Current vs. Temperature (In Circuit of *Figure 2*.)



Accuracy vs. Temperature (Guaranteed)



Accuracy vs. Temperature (Guaranteed)



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Applications

The LM35 can be applied easily in the same way as other integrated-circuit temperature sensors. It can be glued or cemented to a surface and its temperature will be within about 0.01°C of the surface temperature.

This presumes that the ambient air temperature is almost the same as the surface temperature; if the air temperature were much higher or lower than the surface temperature, the actual temperature of the LM35 die would be at an intermediate temperature between the surface temperature and the air temperature. This is expecially true for the TO-92 plastic package, where the copper leads are the principal thermal path to carry heat into the device, so its temperature might be closer to the air temperature than to the surface temperature.

To minimize this problem, be sure that the wiring to the LM35, as it leaves the device, is held at the same temperature as the surface of interest. The easiest way to do this is to cover up these wires with a bead of epoxy which will insure that the leads and wires are all at the same temperature as the surface, and that the LM35 die's temperature will not be affected by the air temperature.

The TO-46 metal package can also be soldered to a metal surface or pipe without damage. Of course, in that case the V− terminal of the circuit will be grounded to that metal. Alternatively, the LM35 can be mounted inside a sealed-end metal tube, and can then be dipped into a bath or screwed into a threaded hole in a tank. As with any IC, the LM35 and accompanying wiring and circuits must be kept insulated and dry, to avoid leakage and corrosion. This is especially true if the circuit may operate at cold temperatures where condensation can occur. Printed-circuit coatings and varnishes such as Humiseal and epoxy paints or dips are often used to insure that moisture cannot corrode the LM35 or its connections.

These devices are sometimes soldered to a small lightweight heat fin, to decrease the thermal time constant and speed up the response in slowly-moving air. On the other hand, a small thermal mass may be added to the sensor, to give the steadiest reading despite small deviations in the air temperature.

Temperature Rise of LM35 Due To Self-heating (Thermal Resistance)

	TO-46, no heat sink	TO-46, small heat fin*	TO-92, no heat sink	TO-92, small heat fin**
Still air	400°C/W	100°C/W	180°C/W	140°C/W
Moving air	100°C/W	40°C/W	90°C/W	70°C/W
Still oil	100°C/W	40°C/W	90°C/W	70°C/W
Stirred oil	50°C/W	30°C/W	45°C/W	40°C/W
(Clamped to metal, Infinite heat sink)	(24°C/W)			

* Wakefield type 201, or 1" disc of 0.020" sheet brass, soldered to case, or similar.

** TO-92 package glued and leads soldered to 1" square of 1/16" printed circuit board with 2 oz. foil or similar.

Typical Applications (Continued)

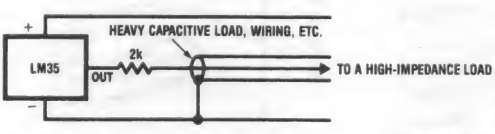

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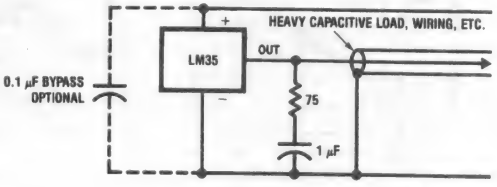
FIGURE 3. LM35 with Decoupling from Capacitive Load



TL/H/5516–20

FIGURE 4. LM35 with R-C Damper

CAPACITIVE LOADS

Like most micropower circuits, the LM35 has a limited ability to drive heavy capacitive loads. The LM35 by itself is able to drive 50 pf without special precautions. If heavier loads are anticipated, it is easy to isolate or decouple the load with a resistor; see *Figure 3*. Or you can improve the tolerance of capacitance with a series R-C damper from output to ground; see *Figure 4*.

When the LM35 is applied with a 200Ω load resistor as shown in *Figure 5, 6,* or *8,* it is relatively immune to wiring capacitance because the capacitance forms a bypass from ground to input, not on the output. However, as with any linear circuit connected to wires in a hostile environment, its performance can be affected adversely by intense electromagnetic sources such as relays, radio transmitters, motors with arcing brushes, SCR transients, etc, as its wiring can act as a receiving antenna and its internal junctions can act as rectifiers. For best results in such cases, a bypass capacitor from V_{IN} to ground and a series R-C damper such as 75Ω in series with 0.2 or 1 μF from output to ground are often useful. These are shown in *Figures 13, 14,* and *16.*

Typical Applications (Continued)

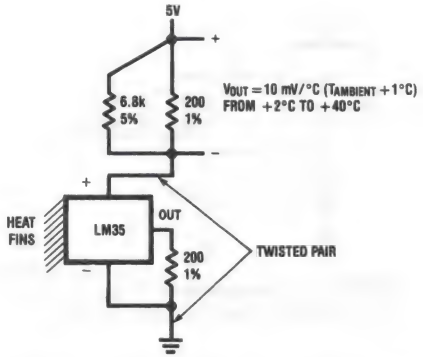


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FIGURE 5. Two-Wire Remote Temperature Sensor (Grounded Sensor)

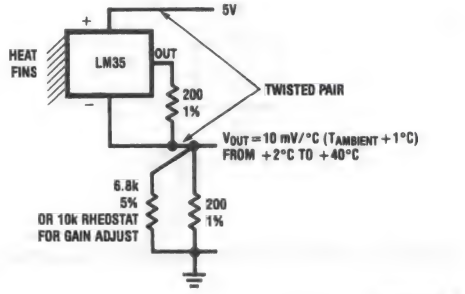

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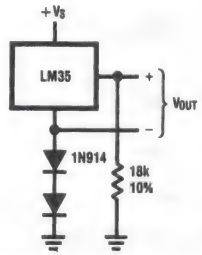
FIGURE 6. Two-Wire Remote Temperature Sensor (Output Referred to Ground)



TL/H/5516–7

FIGURE 7. Temperature Sensor, Single Supply, −55° to +150°C

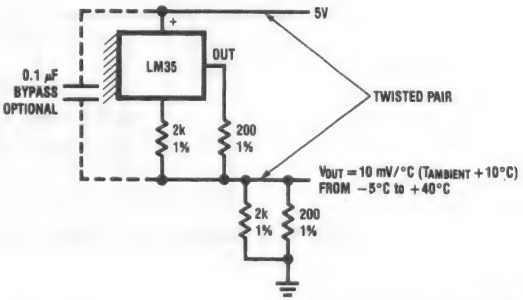


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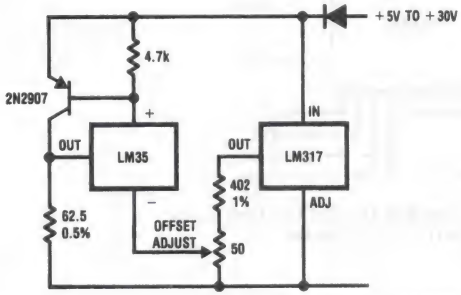
FIGURE 8. Two-Wire Remote Temperature Sensor (Output Referred to Ground)



TL/H/5516–9

FIGURE 9. 4-To-20 mA Current Source (0°C to +100°C)

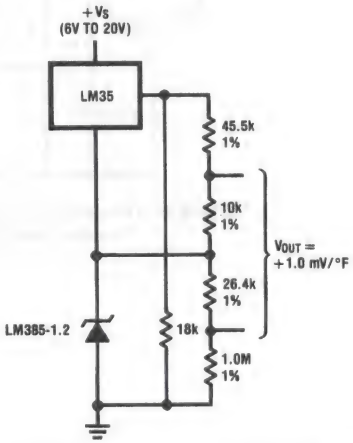


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FIGURE 10. Fahrenheit Thermometer

Typical Applications (Continued)

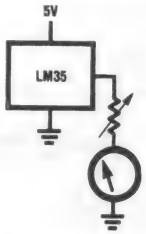

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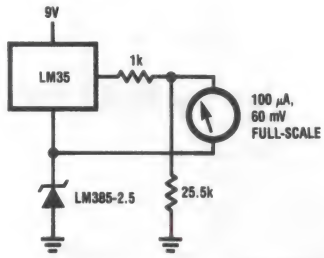
FIGURE 11. Centigrade Thermometer (Analog Meter)



TL/H/5516–12

FIGURE 12. Expanded Scale Thermometer (50° to 80° Fahrenheit, for Example Shown)

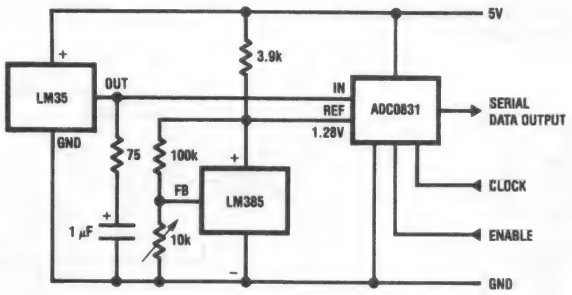


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FIGURE 13. Temperature To Digital Converter (Serial Output) (+ 128°C Full Scale)

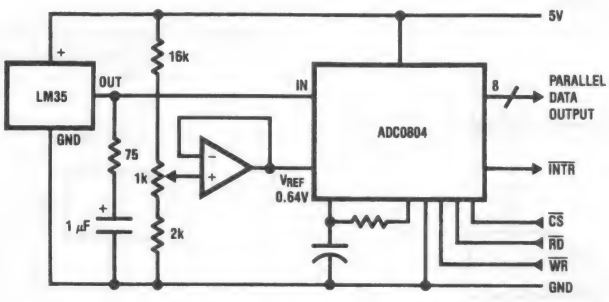

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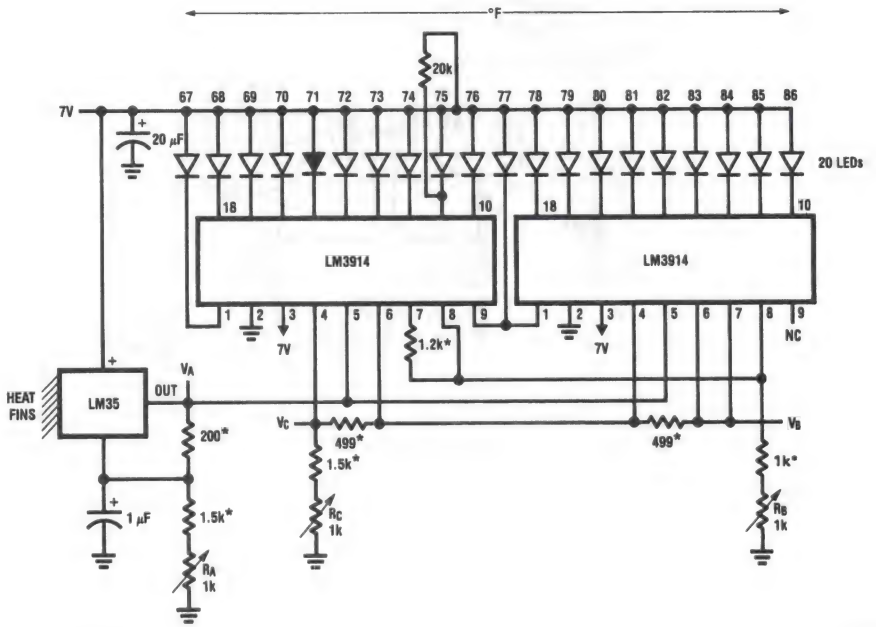
FIGURE 14. Temperature To Digital Converter (Parallel TRI-STATE® Outputs for Standard Data Bus to μP Interface) (128°C Full Scale)

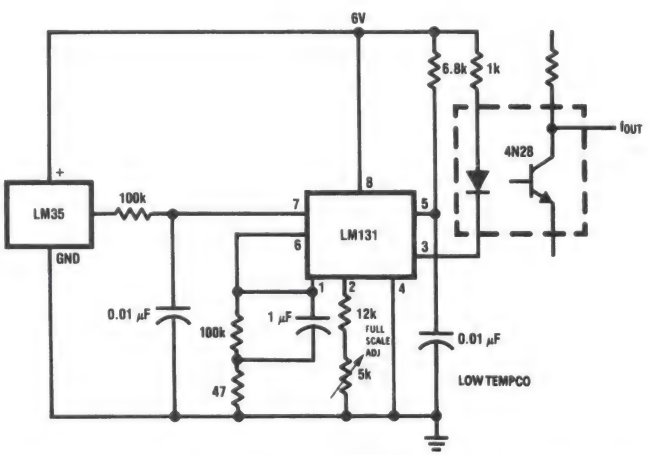
Typical Applications (Continued)



TL/H/5516–16

*=1% or 2% film resistor
-Trim R_B for $V_B=3.075V$
-Trim R_C for $V_C=1.955V$
-Trim R_A for $V_A=0.075V + 100mV/°C \times T_{ambient}$
-Example, $V_A=2.275V$ at 22°C

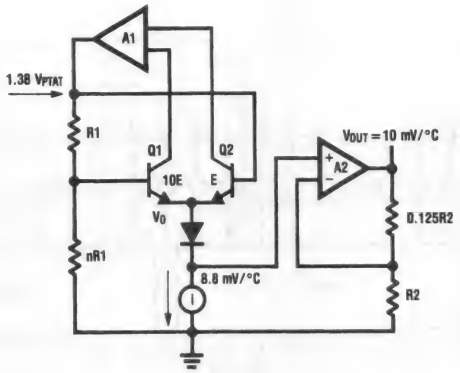
FIGURE 15. Bar-Graph Temperature Display (Dot Mode)



TL/H/5516–15

FIGURE 16. LM35 With Voltage-To-Frequency Converter And Isolated Output (2°C to +150°C; 20 Hz to 1500 Hz)

Block Diagram



TL/H/5516–21

National Semiconductor

LM135/LM235/LM335, LM135A/LM235A/LM335A Precision Temperature Sensors

General Description

The LM135 series are precision, easily-calibrated, integrated circuit temperature sensors. Operating as a 2-terminal zener, the LM135 has a breakdown voltage directly proportional to absolute temperature at +10 mV/°K. With less than 1Ω dynamic impedance the device operates over a current range of 400 μA to 5 mA with virtually no change in performance. When calibrated at 25°C the LM135 has typically less than 1°C error over a 100°C temperature range. Unlike other sensors the LM135 has a linear output.

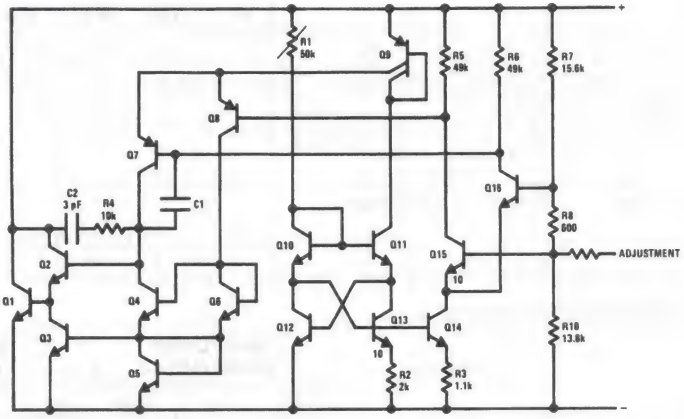
Applications for the LM135 include almost any type of temperature sensing over a −55°C to +150°C temperature range. The low impedance and linear output make interfacing to readout or control circuitry especially easy.

The LM135 operates over a −55°C to +150°C temperature range while the LM235 operates over a −40°C to +125°C temperature range. The LM335 operates from −40°C to +100°C. The LM135/LM235/LM335 are available packaged in hermetic TO-46 transistor packages while the LM335 is also available in plastic TO-92 packages.

Features

- Directly calibrated in °Kelvin
- 1°C initial accuracy available
- Operates from 400 μA to 5 mA
- Less than 1Ω dynamic impedance
- Easily calibrated
- Wide operating temperature range
- 200°C overrange
- Low cost

Schematic Diagram



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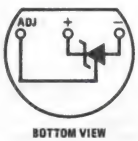
Connection Diagrams

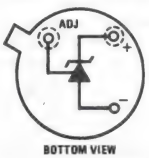
TO-92
Plastic Package



BOTTOM VIEW

Order Number LM335Z or LM335AZ
See NS Package Number Z03A

TO-46
Metal Can Package*



BOTTOM VIEW

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*Case is connected to negative pin

Order Number LM135H, LM235H, LM335H, LM135AH, LM235AH or LM335AH
See NS Package Number H03H

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

(Note 4)

Reverse Current	15 mA
Forward Current	10 mA
Storage Temperature	
TO-46 Package	−60°C to +180°C
TO-92 Package	−60°C to +150°C

Specified Operating Temp. Range

	Continuous	Intermittent (Note 2)
LM135, LM135A	−55°C to +150°C	150°C to 200°C
LM235, LM235A	−40°C to +125°C	125°C to 150°C
LM335, LM335A	−40°C to +100°C	100°C to 125°C

Lead Temp. (Soldering, 10 seconds)

TO-92 Package:	260°C
TO-46 Package:	300°C

Temperature Accuracy LM135/LM235, LM135A/LM235A (Note 1)

Parameter	Conditions	LM135A/LM235A			LM135/LM235			Units
		Min	Typ	Max	Min	Typ	Max	
Operating Output Voltage	$T_C = 25°C$, $I_R = 1$ mA	2.97	2.98	2.99	2.95	2.98	3.01	V
Uncalibrated Temperature Error	$T_C = 25°C$, $I_R = 1$ mA		0.5	1		1	3	°C
Uncalibrated Temperature Error	$T_{MIN} \le T_C \le T_{MAX}$, $I_R = 1$ mA		1.3	2.7		2	5	°C
Temperature Error with 25°C Calibration	$T_{MIN} \le T_C \le T_{MAX}$, $I_R = 1$ mA		0.3	1		0.5	1.5	°C
Calibrated Error at Extended Temperatures	$T_C = T_{MAX}$ (Intermittent)		2			2		°C
Non-Linearity	$I_R = 1$ mA		0.3	0.5		0.3	1	°C

Temperature Accuracy LM335, LM335A (Note 1)

Parameter	Conditions	LM335A			LM335			Units
		Min	Typ	Max	Min	Typ	Max	
Operating Output Voltage	$T_C = 25°C$, $I_R = 1$ mA	2.95	2.98	3.01	2.92	2.98	3.04	V
Uncalibrated Temperature Error	$T_C = 25°C$, $I_R = 1$ mA		1	3		2	6	°C
Uncalibrated Temperature Error	$T_{MIN} \le T_C \le T_{MAX}$, $I_R = 1$ mA		2	5		4	9	°C
Temperature Error with 25°C Calibration	$T_{MIN} \le T_C \le T_{MAX}$, $I_R = 1$ mA		0.5	1		1	2	°C
Calibrated Error at Extended Temperatures	$T_C = T_{MAX}$ (Intermittent)		2			2		°C
Non-Linearity	$I_R = 1$ mA		0.3	1.5		0.3	1.5	°C

Electrical Characteristics (Note 1)

Parameter	Conditions	LM135/LM235 LM135A/LM235A			LM335 LM335A			Units
		Min	Typ	Max	Min	Typ	Max	
Operating Output Voltage Change with Current	400 μA ≤ I_R ≤ 5 mA At Constant Temperature		2.5	10		3	14	mV
Dynamic Impedance	$I_R = 1$ mA		0.5			0.6		Ω
Output Voltage Temperature Coefficient			+10			+10		mV/°C
Time Constant	Still Air 100 ft/Min Air Stirred Oil		80 10 1			80 10 1		sec sec sec
Time Stability	$T_C = 125°C$		0.2			0.2		°C/khr

Note 1: Accuracy measurements are made in a well-stirred oil bath. For other conditions, self heating must be considered.

Note 2: Continuous operation at these temperatures for 10,000 hours for H package and 5,000 hours for Z package may decrease life expectancy of the device.

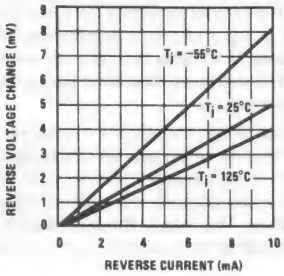
Note 3: Thermal Resistance

	TO-92	TO-46
θ_{JA} (junction to ambient)	202°C/W	400°C/W
θ_{JC} (junction to case)	170°C/W	N/A

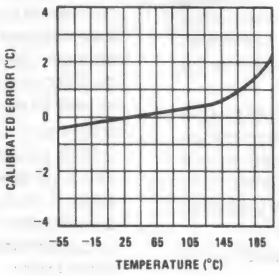
Note 4: Refer to RETS135H for military specifications.

Typical Performance Characteristics

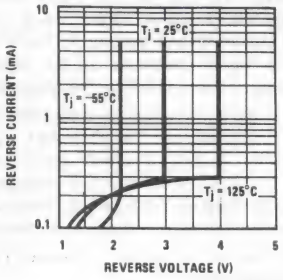
Reverse Voltage Change



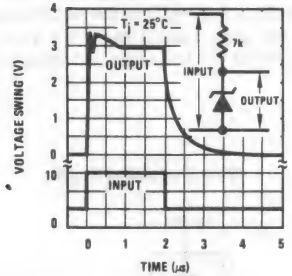
Calibrated Error



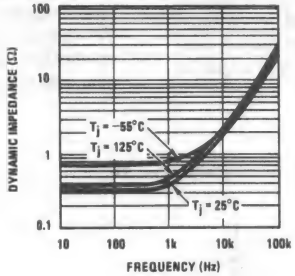
Reverse Characteristics



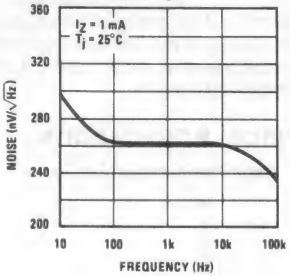
Response Time



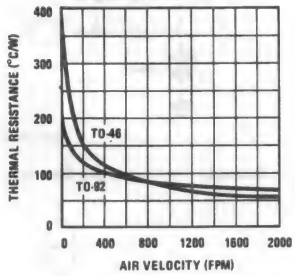
Dynamic Impedance



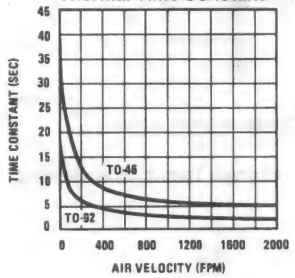
Noise Voltage



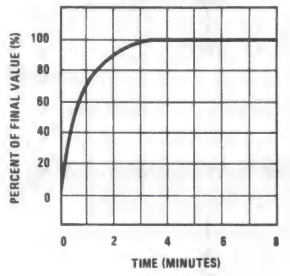
Thermal Resistance Junction to Air



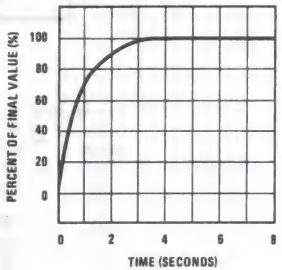
Thermal Time Constant



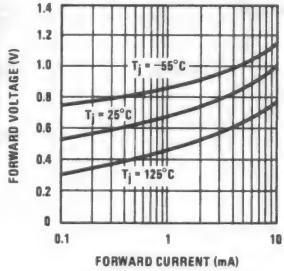
Thermal Response in Still Air



Thermal Response in Stirred Oil Bath



Forward Characteristics



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Application Hints

CALIBRATING THE LM135

Included on the LM135 chip is an easy method of calibrating the device for higher accuracies. A pot connected across the LM135 with the arm tied to the adjustment terminal allows a 1-point calibration of the sensor that corrects for inaccuracy over the full temperature range.

This single point calibration works because the output of the LM135 is proportional to absolute temperature with the extrapolated output of sensor going to 0V output at 0°K (−273. 15°C). Errors in output voltage versus temperature are only slope (or scale factor) errors so a slope calibration at one temperature corrects at all temperatures.

The output of the device (calibrated or uncalibrated) can be expressed as:

$$V_{OUT_T} = V_{OUT_{T_o}} \times \frac{T}{T_o}$$

where T is the unknown temperature and T_o is a reference temperature, both expressed in degrees Kelvin. By calibrating the output to read correctly at one temperature the output at all temperatures is correct. Nominally the output is calibrated at 10 mV/°K.

To insure good sensing accuracy several precautions must be taken. Like any temperature sensing device, self heating can reduce accuracy. The LM135 should be operated at the lowest current suitable for the application. Sufficient current, of course, must be available to drive both the sensor and the calibration pot at the maximum operating temperature as well as any external loads.

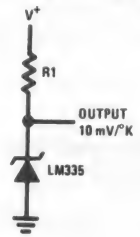
If the sensor is used in an ambient where the thermal resistance is constant, self heating errors can be calibrated out. This is possible if the device is run with a temperature stable current. Heating will then be proportional to zener voltage and therefore temperature. This makes the self heating error proportional to absolute temperature the same as scale factor errors.

WATERPROOFING SENSORS

Meltable inner core heat shrinkable tubing such as manufactured by Raychem can be used to make low-cost waterproof sensors. The LM335 is inserted into the tubing about ½" from the end and the tubing heated above the melting point of the core. The unfilled ½" end melts and provides a seal over the device.

Typical Applications

Basic Temperature Sensor



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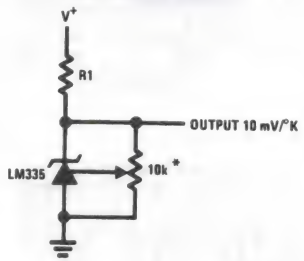
Calibrated Sensor



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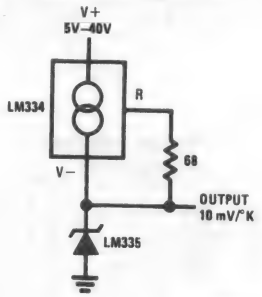
*Calibrate for 2.982V at 25°C

Wide Operating Supply



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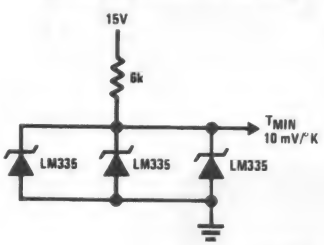
Minimum Temperature Sensing



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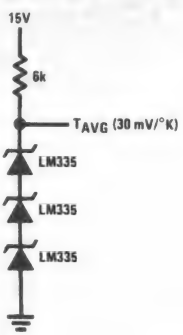
Average Temperature Sensing



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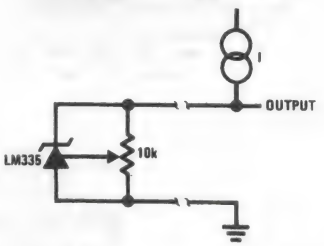
Remote Temperature Sensing



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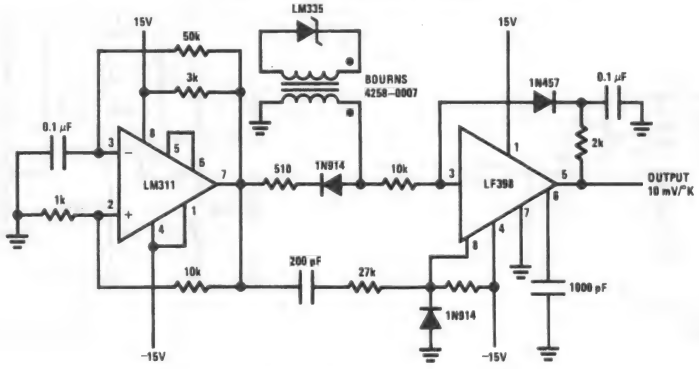
Wire length for 1°C error due to wire drop

AWG	I_R = 1 mA FEET	I_R = 0.5 mA* FEET
14	4000	8000
16	2500	5000
18	1600	3200
20	1000	2000
22	625	1250
24	400	800

*For I_R = 0.5 mA, the trim pot must be deleted.

Typical Applications (Continued)

Isolated Temperature Sensor



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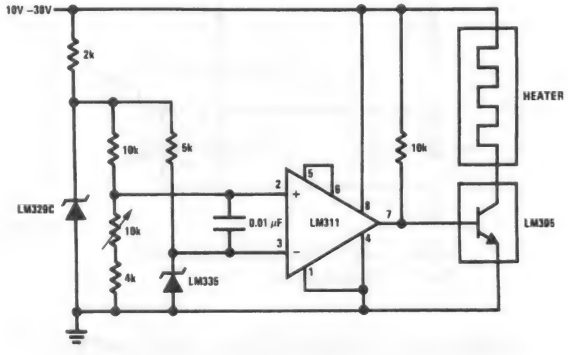
Simple Temperature Controller



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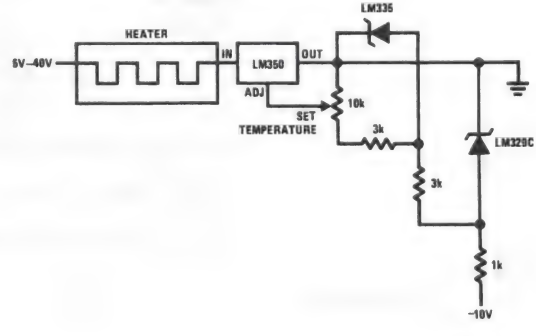
Simple Temperature Control



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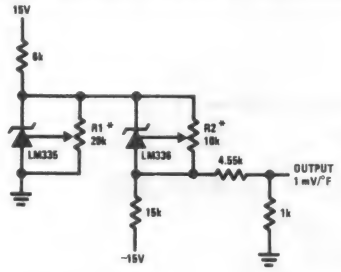
Typical Applications (Continued)

Ground Referred Fahrenheit Thermometer



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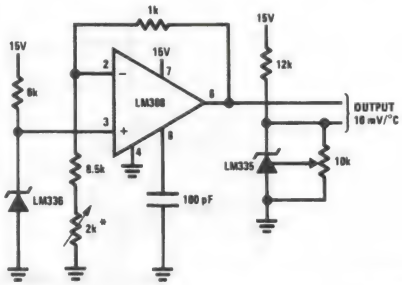
*Adjust R2 for 2.554V across LM336.
Adjust R1 for correct output.

Centigrade Thermometer



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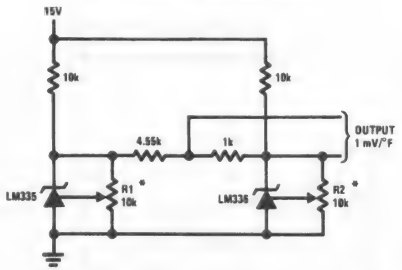
*Adjust for 2.7315V at output of LM308

Fahrenheit Thermometer



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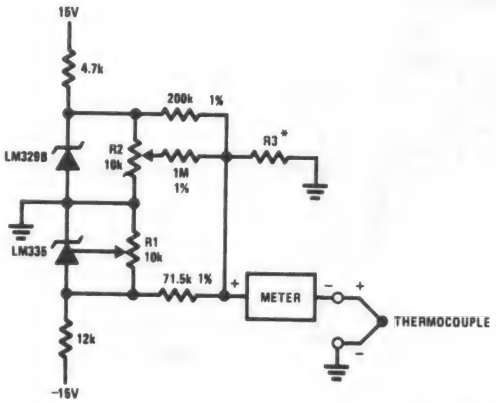
*To calibrate adjust R2 for 2.554V across LM336.
Adjust R1 for correct output.

THERMOCOUPLE COLD JUNCTION COMPENSATION
Compensation for Grounded Thermocouple



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*Select R3 for proper thermocouple type

THERMOCOUPLE	R3 (±1%)	SEEBECK COEFFICIENT
J	377Ω	52.3 μV/°C
T	308Ω	42.8 μV/°C
K	293Ω	40.8 μV/°C
S	45.8Ω	6.4 μV/°C

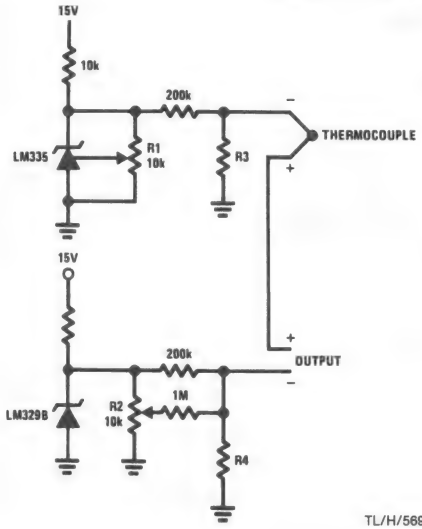
Adjustments: Compensates for both sensor and resistor tolerances

1. Short LM329B
2. Adjust R1 for Seebeck Coefficient times ambient temperature (in degrees K) across R3.
3. Short LM335 and adjust R2 for voltage across R3 corresponding to thermocouple type

J	14.32 mV	K	11.17 mV
T	11.79 mV	S	1.768 mV

Typical Applications (Continued)

Single Power Supply Cold Junction Compensation



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*Select R3 and R4 for thermocouple type

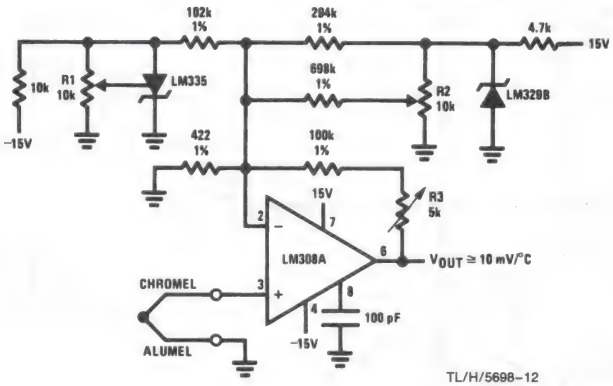
THERMO-COUPLE	R3	R4	SEEBECK COEFFICIENT
J	1.05K	385Ω	52.3 μV/°C
T	856Ω	315Ω	42.8 μV/°C
K	816Ω	300Ω	40.8 μV/°C
S	128Ω	46.3Ω	6.4 μV/°C

Adjustments:

1. Adjust R1 for the voltage across R3 equal to the Seebeck Coefficient times ambient temperature in degrees Kelvin.
2. Adjust R2 for voltage across R4 corresponding to thermocouple

J	14.32 mV
T	11.79 mV
K	11.17 mV
S	1.768 mV

Centigrade Calibrated Thermocouple Thermometer



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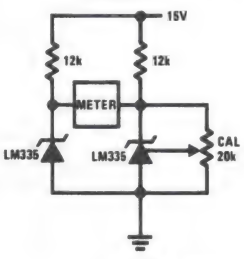
Terminate thermocouple reference junction in close proximity to LM335.

Adjustments:

1. Apply signal in place of thermocouple and adjust R3 for a gain of 245.7.
2. Short non-inverting input of LM308A and output of LM329B to ground.
3. Adjust R1 so that V_{OUT} = 2.982V @ 25°C.
4. Remove short across LM329B and adjust R2 so that V_{OUT} = 246 mV @ 25°C.
5. Remove short across thermocouple.

Differential Temperature Sensor



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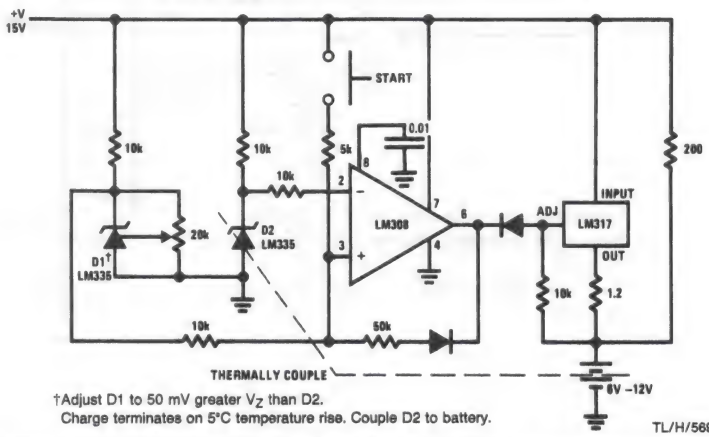
Fast Charger for Nickel-Cadmium Batteries



†Adjust D1 to 50 mV greater V_Z than D2.
Charge terminates on 5°C temperature rise. Couple D2 to battery.

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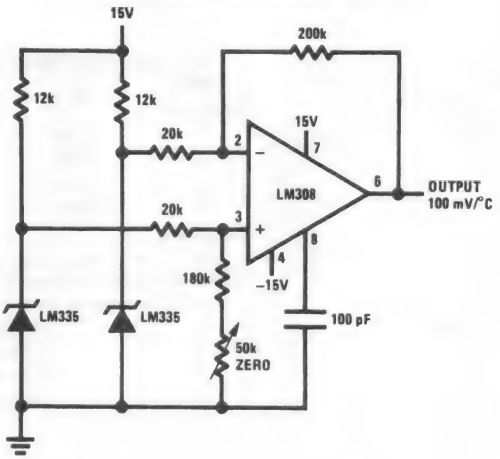
Typical Applications (Continued)

Differential Temperature Sensor



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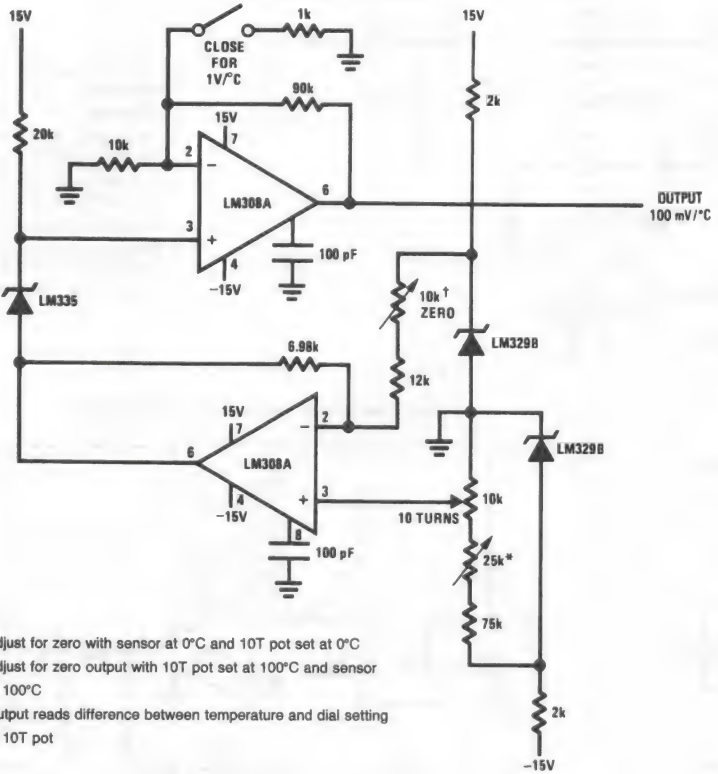
Variable Offset Thermometer‡



†Adjust for zero with sensor at 0°C and 10T pot set at 0°C

*Adjust for zero output with 10T pot set at 100°C and sensor at 100°C

‡Output reads difference between temperature and dial setting of 10T pot

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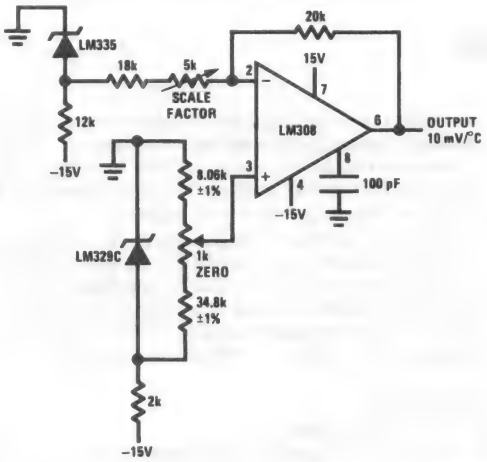
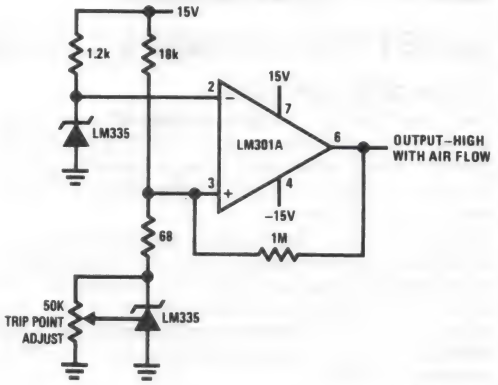
Typical Applications (Continued)

Ground Referred Centigrade Thermometer



TL/H/5698–16

Air Flow Detector*



TL/H/5698–17

*Self heating is used to detect air flow

Definition of Terms

Operating Output Voltage: The voltage appearing across the positive and negative terminals of the device at specified conditions of operating temperature and current.

Uncalibrated Temperature Error: The error between the operating output voltage at 10 mV/°K and case temperature at specified conditions of current and case temperature.

Calibrated Temperature Error: The error between operating output voltage and case temperature at 10 mV/°K over a temperature range at a specified operating current with the 25°C error adjusted to zero.

National Semiconductor

LM3911 Temperature Controller

General Description

The LM3911 is a highly accurate temperature measurement and/or control system for use over a −25°C to +85°C temperature range. Fabricated on a single monolithic chip, it includes a temperature sensor, a stable voltage reference and an operational amplifier.

The output voltage of the LM3911 is directly proportional to temperature in degrees Kelvin at 10 mV/°K. Using the internal op amp with external resistors any temperature scale factor is easily obtained. By connecting the op amp as a comparator, the output will switch as the temperature transverses the set-point making the device useful as an on-off temperature controller.

An active shunt regulator is connected across the power leads of the LM3911 to provide a stable 6.8V voltage reference for the sensing system. This allows the use of any power supply voltage with suitable external resistors.

The input bias current is low and relatively constant with temperature, ensuring high accuracy when high source impedance is used. Further, the output collector can be returned to a voltage higher than 6.8V allowing the LM3911 to drive lamps and relays up to a 35V supply.

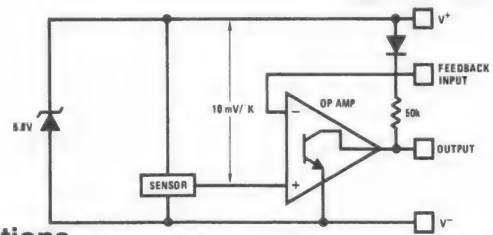
The LM3911 uses the difference in emitter-base voltage of transistors operating at different current densities as the basic temperature sensitive element. Since this output depends only on transistor matching the same reliability and stability as present op amps can be expected.

The LM3911 is available in two package styles, a metal can TO-46 and an 8-lead epoxy mini-DIP. In the epoxy package all electrical connections are made on one side of the device allowing the other 4 leads to be used for attaching the LM3911 to the temperature souce. The LM3911 is rated for operation over a −25°C to +85°C temperature range.

Features

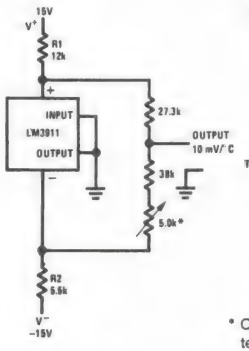
- Uncalibrated accuracy ±10°C
- Internal op amp with frequency compensation
- Linear output of 10 mV/°K (10 mV/°C)
- Can be calibrated in degrees Kelvin, Celsius or Fahrenheit
- Output can drive loads up to 35V
- Internal stable voltage reference
- Low cost

Block Diagram



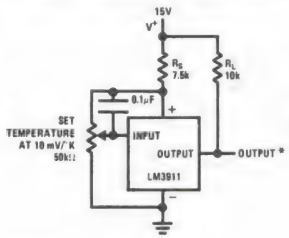
Typical Applications

Ground Referred Centigrade Thermometer



* Trims out initial zener tolerance. Set output to read C

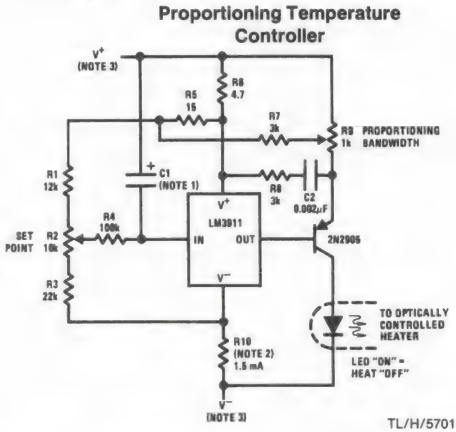
Basic Temperature Controller



* Output goes negative on temperature increase

$R_S = (V^+ - 6.8V)\ k\Omega$

Proportioning Temperature Controller



TL/H/5701–1

Note 1: C1 determines proportioning frequency $f \approx \frac{1}{2R4\ C1}$

Note 2: $R10 = \frac{|V+| + |V-| - 7V}{0.0015A}$

Note 3: Either V− or V+ can be ground.

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Current (Externally Set)	10 mA
Output Collector Voltage, V++	36V
Feedback Input Voltage Range	0V to +7.0V
Output Short Circuit Duration	Indefinite
Operating Temperature Range	−25°C to +85°C
Storage Temperature Range	−65°C to +150°C
Lead Temperature (Soldering, 10 seconds)	260°C

Electrical Characteristics (Note 1)

Parameter	Conditions	Min	Typ	Max	Units
SENSOR					
Output Voltage	$T_A = -25°C$, (Note 2)	2.36	2.48	2.60	V
Output Voltage	$T_A = +25°C$, (Note 2)	2.88	2.98	3.08	V
Output Voltage	$T_A = +85°C$, (Note 2)	3.46	3.58	3.70	V
Linearity	$\Delta T = 100°C$		0.5	2	%
Long-Term Stability			0.3		%
Repeatability			0.3		%
VOLTAGE REFERENCE					
Reverse Breakdown Voltage	$1\ mA \le I_Z \le 5\ mA$	6.55	6.85	7.25	V
Reverse Breakdown Voltage Change With Current	$1\ mA \le I_Z \le 5\ mA$		10	35	mV
Temperature Stability			20	85	mV
Dynamic Impedance	$I_Z = 1\ mA$		3.0		Ω
RMS Noise Voltage	$10\ Hz \le f \le 10\ kHz$		30		μV
Long Term Stability	$T_A = +85°C$		6.0		mV
OP AMP					
Input Bias Current	$T_A = +25°C$		35	150	nA
Input Bias Current			45	250	nA
Voltage Gain	$R_L = 36k$, V++ = 36V	2500	15000		V/V
Output Leakage Current	$T_A = 25°C$ (Note 3)		0.2	2	μA
Output Leakage Current	(Note 3)		1.0	8	μA
Output Source Current	$V_{OUT} \le 3.70$	10			μA
Output Sink Current	$1V \le V_{OUT} \le 36V$	2.0			mA

Note 1: These specifications apply for $-25°C \le T_A \le +85°C$ and $0.9\ mA \le I_{SUPPLY} \le 1.1\ mA$ unless otherwise specified; $C_L \le 50\ pF$.

Note 2: The output voltage applies to the basic thermometer configuration with the output and input terminals shorted and a load resistance of ≥ 1.0 MΩ. This is the feedback sense voltage and includes errors in both the sensor and op amp. This voltage is specified for the sensor in a rapidly stirred oil bath. The output is referred to V+.

Note 3: The output leakage current is specified with ≥ 100 mV overdrive. Since this voltage changes with temperature, the voltage drive for turn-off changes and is defined as V_{OUT} (with output and input shorted) −100 mV. This specification applies for $V_{OUT} = 36V$.

Application Hints

Although the LM3911 is designed to be totally trouble-free, certain precautions should be taken to insure the best possible performance.

As with any temperature sensor, internal power dissipation will raise the sensor's temperature above ambient. Nominal suggested operating current for the shunt regulator is 1.0 mA and causes 7.0 mW of power dissipation. In free, still, air this raises the package temperature by about 1.2°K. Although the regulator will operate at higher reverse currents and the output will drive loads up to 5.0 mA, these higher currents will raise the sensor temperature to about 19°K above ambient-degrading accuracy. Therefore, the sensor should be operated at the lowest possible power level.

With moving air, liquid or surface temperature sensing, self-heating is not as great a problem since the measured media will conduct the heat from the sensor. Also, there are many small heat sinks designed for transistors which will improve heat transfer to the sensor from the surrounding medium. A small finned clip-on heat sink is quite effective in free-air. It should be mentioned that the LM3911 die is on the base of the package and therefore coupling to the base is preferable.

The internal reference regulator provides a temperature stable voltage for offsetting the output or setting a comparison point in temperature controllers. However, since this reference is at the same temperature as the sensor temperature, changes will also cause reference drift. For application where maximum accuracy is needed an external reference should be used. Of course, for fixed temperature controllers the internal reference is adequate.

Typical Performance Characteristics

Temperature Conversion

$T_{CENTIGRADE} = T_C$

$T_{FAHRENHEIT} = T_F$

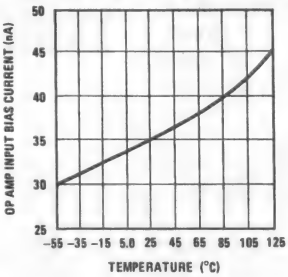
$T_{KELVIN} = T_K$

$T_K = T_C + 273.16$

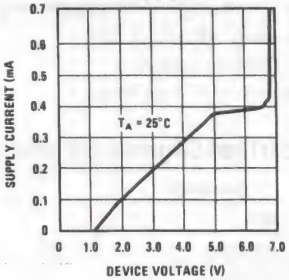
$T_C = (40 + T_F)\frac{5}{9} - 40$

$T_F = (40 + T_C)\frac{9}{5} - 40$

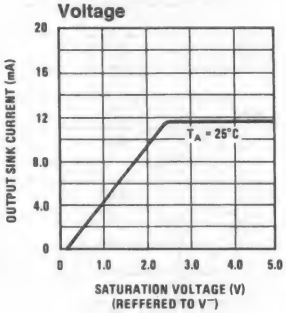
Op Amp Input Current



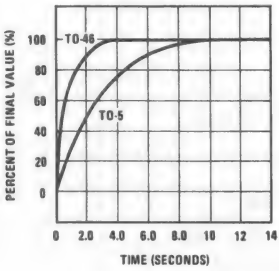
Power Supply Current



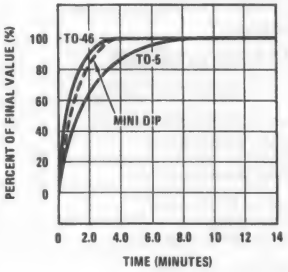
Output Saturation Voltage



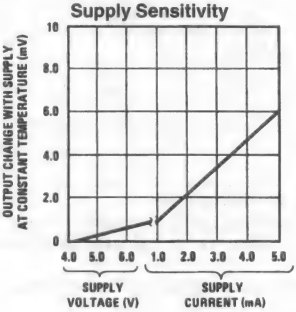
Thermal Time Constant in Stirred Oil Bath



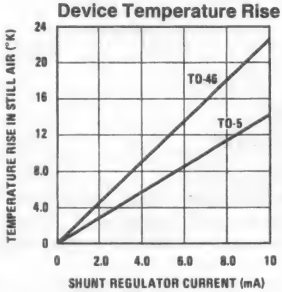
Thermal Time Constant in Still Air



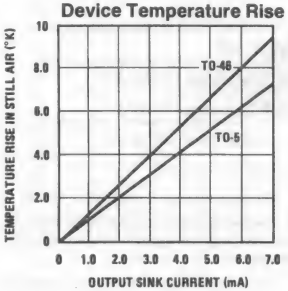
Supply Sensitivity



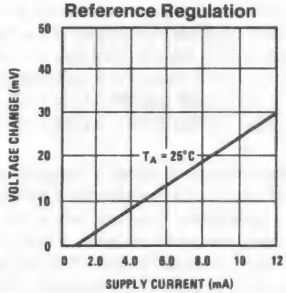
Device Temperature Rise



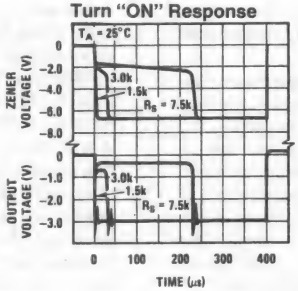
Device Temperature Rise



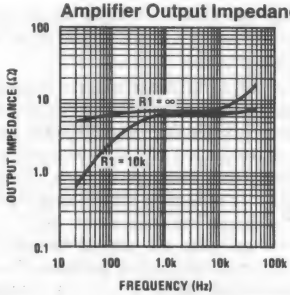
Reference Regulation



Turn "ON" Response

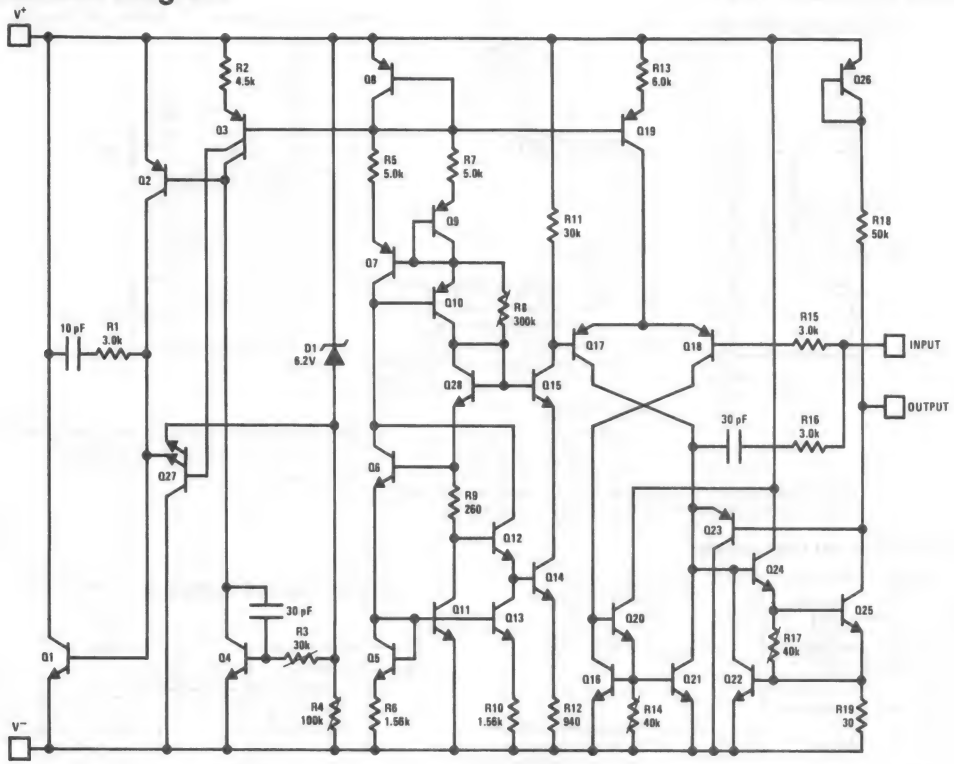


Amplifier Output Impedance



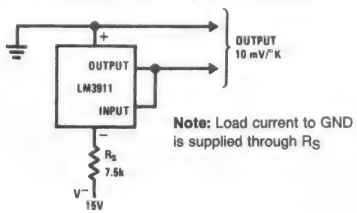
TL/H/5701–2

Schematic Diagram



Typical Applications (Continued)

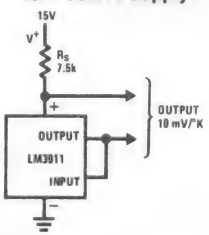
Basic Thermometer for Negative Supply



Note: Load current to GND is supplied through R_S

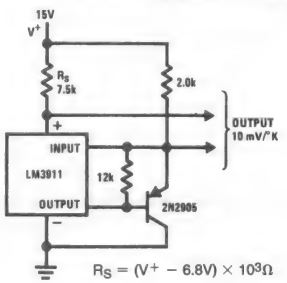
$R_S = (V^- - 6.8V) \times 10^3\Omega$

Basic Thermometer for Positive Supply



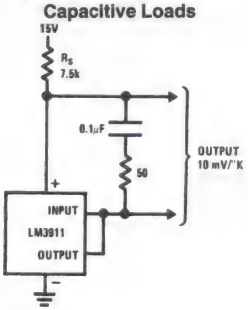
$R_S = (V^- - 6.8V) \times 10^3\Omega$

Increasing Output Drive

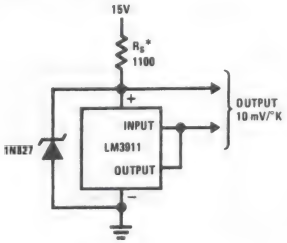


$R_S = (V^+ - 6.8V) \times 10^3\Omega$

External Frequency Compensation for Greater Stability when Driving Capacitive Loads

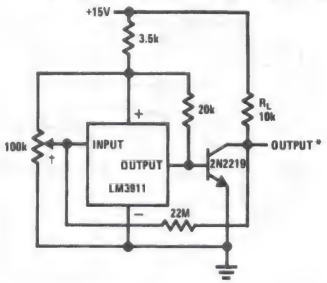


Operating With External Zener for Lower Power Dissipation



*Depends on Zener current.

Temperature Controller With Hysteresis



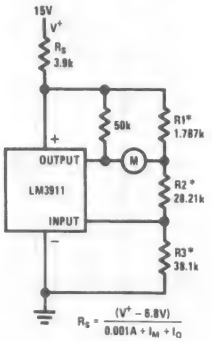
*Output goes positive on temperature increase

†Set temperature

TL/H/5701–3

Typical Applications (Continued)

Thermometer With Meter Output



$$R1^* = \frac{(V_Z)\,0.01\Delta T}{I_M\,(V_Z - 0.01\,T_O)} \quad **$$

$$\text{Select } I_Q \leq \frac{2V}{R1}$$

$$R2 = \frac{0.01\,T_O - I_Q R1}{I_Q}$$

$$R3 = \frac{V_Z}{I_Q} - R1 - R2$$

$$\left(I_Q \leq \frac{2V}{R1}\right)$$

V_Z = Shunt regulator voltage (use 6.85)
ΔT = Meter temperature span (°K)
I_M = Meter full scale current (A)
T_O = Meter zero temperature (°K)
I_Q = Current through R1, R2, R3 at zero meter current (10 μA to 1.0 mA) (A)

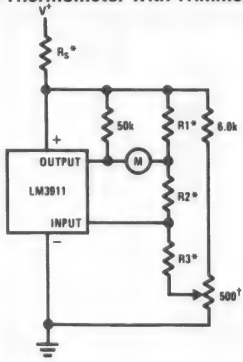
*Values shown for:

$T_O = 300°K$, $\Delta T = 100°K$,

$I_M = 1.0$ mA, $I_Q = 100$ μA

**The 0.01 in the above and following equations is in units of V/°K or V/°C, and is a result of the basic 0.01V/°K sensitivity of the transducer

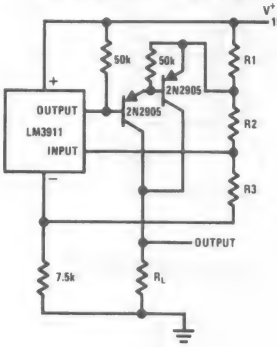
Meter Thermometer With Trimmed Output



*Selected as for meter thermometer except T_O should be 5°K more than desired and $I_Q = 100$ μA

†Calibrates T_O

Ground Referred Thermometer



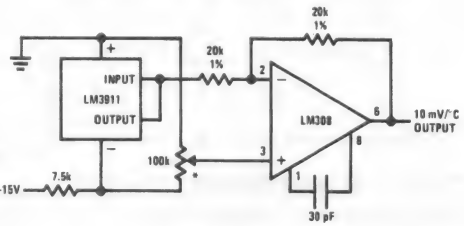
$$R1 = \frac{(V_Z)(10mV)(\Delta T)}{\frac{V_O}{R_L}\,(V_Z - 0.01\,T_O)}$$

$$R2 = \frac{0.01\,T_O - I_Q R1}{I_Q}$$

$$R3 = \frac{V_Z}{I_Q} - R1 - R2$$

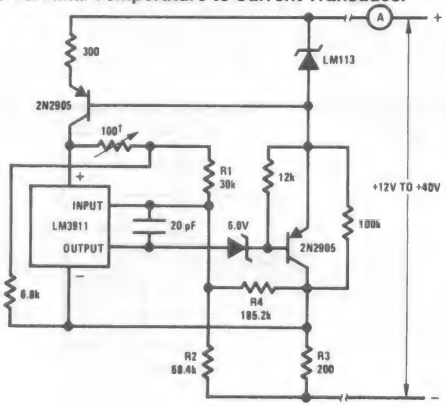
V_Z = Shunt regulator voltage
ΔT = Temperature span (°K)
T_O = Temperature for zero output (°K)
V_O = Full scale output voltage ≤ 10V
I_Q = Current through R1, R2, R3 at zero output voltage (typically 100 μA to 1.0 mA)

Ground Referred Centigrade Thermometer



*Set zero

Two Terminal Temperature to Current Transducer*



$$R2\,(\Omega) = \frac{\left(V_Z - 0.01\,T_L\right)\left(I_H - \frac{0.01\,T_H}{R1}\right) + \left(V_Z - 0.01\,T_H\right)\left(\frac{0.01\,T_L}{R1} - I_L\right)}{\frac{0.01}{R1\,R3}\left[T_H(V_Z - 0.01\,T_L) - T_L(V_Z - 0.01\,T_H)\right]} \quad **$$

$$R3(\Omega) \geq \frac{V_Z\left(\frac{T_H}{T_L} - 1\right)}{I_H - \frac{I_L T_H}{T_L}}$$

$$\frac{1}{R4} = \frac{1}{(V_Z - 0.01\,T_L)(R2)}\left[\frac{(R2)(0.01\,T_L)}{R1} + \frac{\left(\frac{V_Z - 0.01\,T_L}{R2} - I_L\right)}{\frac{1}{R2} + \frac{1}{R3}}\right] - \frac{1}{R2}$$

T_L = Temperature for I_L (K)
T_H = Temperature for I_H (K)
V_Z = Zener voltage (V)
I_L = Low temperature output current (A)
I_H = High temperature output current (A)

*Values shown for $I_{OUT} = 1$ mA to 10 mA for 10°F to 100°F

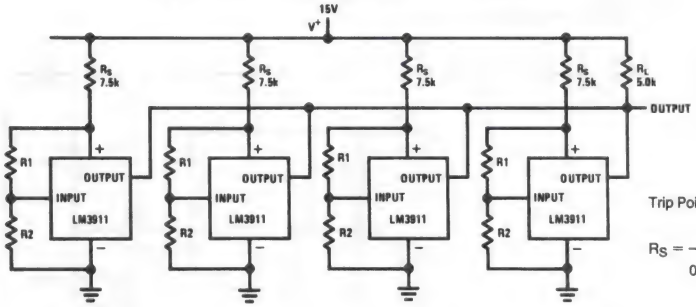
†Set temperature

TL/H/5701–4

**The 0.01 in the above and following equations is in units of V/°K or V/°C, and is a result of the basic 0.01V/°K sensitivity of the transducer

Typical Applications (Continued)

Over Temperature Detectors With Common Output

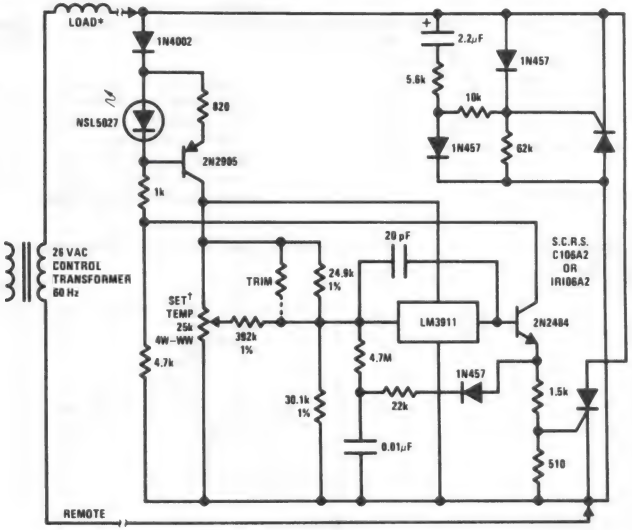


$$\text{Trip Point} = V_Z \frac{R1}{R1 + R2}$$

$$R_S = \frac{(V^+ - 6.8V)}{0.001\ A + \frac{6.8V}{R1 + R2}}$$

TL/H/5701–5

Two-Wire Remote A.C. Electronic Thermostat (Gas or Oil Furnace Control)



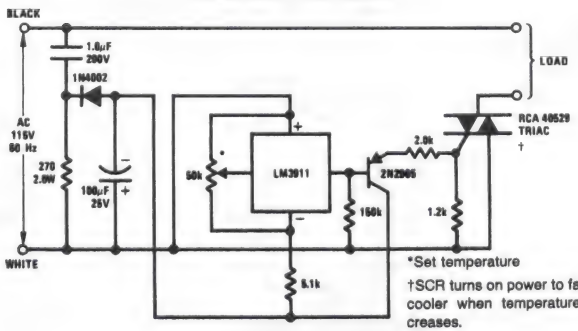
TL/H/5701–8

*Solenoid or 6–15W heater

†Pot will provide about a 50°F to 90°F setting range. The trim resistor (100k) is selected to bring 70°F near the middle of the pot rotation.

SCR heating, by proper positioning, can preheat the sensor giving control anticipation as is presently used in many home thermostats.

Electronic Thermostat



*Set temperature

†SCR turns on power to fan or cooler when temperature increases.

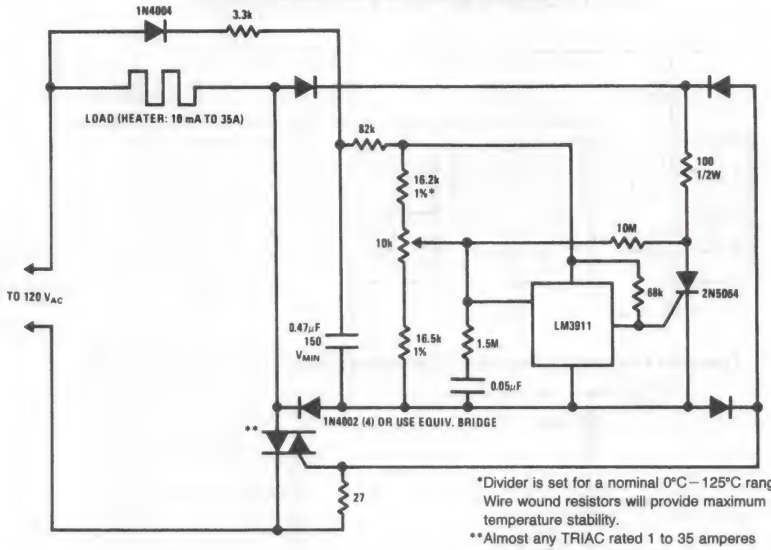
TL/H/5701–9

SENSITIVE GATE TRIAC
$I_{GT} = \leq 5$ mA
RCA T2300, 40529
OR SIMILAR

6

Typical Applications (Continued)

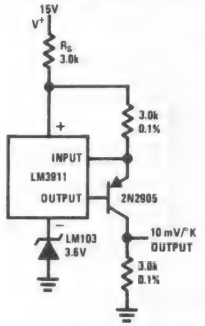
Three-Wire Electronic Thermostat



*Divider is set for a nominal 0°C–125°C range. Wire wound resistors will provide maximum temperature stability.

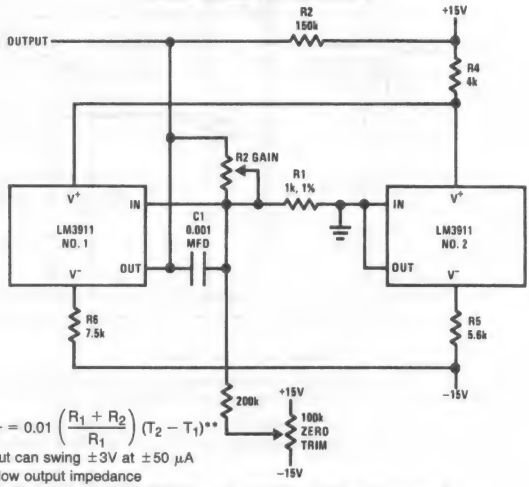
**Almost any TRIAC rated 1 to 35 amperes usable with appropriate load.

Kelvin Thermometer With Ground Referred Output



$$R_S = \frac{V_S^+ - 6.8V \times 10^3\Omega}{2}$$

Differential Thermometer



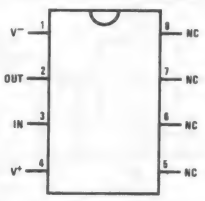
$$V_{OUT} = 0.01\left(\frac{R_1 + R_2}{R_1}\right)(T_2 - T_1)^{**}$$

Output can swing ±3V at ±50 μA with low output impedance

**The 0.01 in the above equation is in units of V/°K or V/°C, and is a result of the basic 0.01 V/°K sensitivity of the transducer

Connection Diagrams

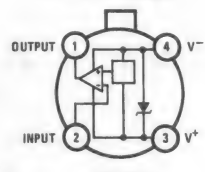
Dual-In-Line Package



TL/H/5701–6

Order Number LM3911N
See NS Package N08E

TO-46 Package



Note: Pin 4 connected to case.

TL/H/5701–7

Order Number LM3911H-46
See NS Package H04A

Section 7

Voltage References

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National Semiconductor

Voltage Reference Selection Guide

Shunt Type

Reverse Breakdown Voltage (V_R)	Device	Operating Temp. Range*	Voltage Tolerance Max, $T_A = 25°C$	Temperature Drift		Operating Current Range, I_R	Output Dynamic Impedance (Typ)
				ppm/°C (Max)	Over Range		
1.22	LM113-2	M	±1%	50 (Typ)	−55°C to +125°C	500 μA to 20 mA	0.8
1.22	LM113-1	M	±2%	50 (Typ)	−55°C to +125°C	500 μA to 20 mA	0.8
1.22	LM113	M	±5%	100 (Typ)	−55°C to +125°C	500 μA to 20 mA	0.8
1.22	LM313	C	±5%	100 (Typ)	0°C to +70°C	500 μA to 20 mA	0.8
1.235	LM185BX-1.2	M	±1%	30	−55°C to +125°C	10 μA to 20 mA	1
1.235	LM185BY-1.2	M	±1%	50	−55°C to +125°C	10 μA to 20 mA	1
1.235	LM185-1.2	M	±1%	150	−55°C to +125°C	10 μA to 20 mA	1
1.235	LM285BX-1.2	I	±1%	30	−40°C to +85°C	10 μA to 20 mA	1
1.235	LM285BY-1.2	I	±1%	50	−40°C to +85°C	10 μA to 20 mA	1
1.235	LM285-1.2	I	±1%	150	−40°C to +85°C	10 μA to 20 mA	1
1.235	LM385BX-1.2	C	±1%	30	0°C to +70°C	15 μA to 20 mA	1
1.235	LM385BY-1.2	C	±1%	50	0°C to +70°C	15 μA to 20 mA	1
1.235	LM385B-1.2	C	±1%	150	0°C to +70°C	15 μA to 20 mA	1
1.235	LM385-1.2	C	+2%, −2.4%	150	0°C to +70°C	15 μA to 20 mA	1
1.24 to 5.3 (Adj.)	LM185B	M	±1%	150	−55°C to +125°C	10 μA to 20 mA	0.3
1.24 to 5.3 (Adj.)	LM185BX	M	±1%	30	−55°C to +125°C	10 μA to 20 mA	0.3
1.24 to 5.3 (Adj.)	LM185BY	M	±1%	50	−55°C to +125°C	10 μA to 20 mA	0.3
1.24 to 5.3 (Adj.)	LM285BX	I	±1%	30	−40°C to +85°C	10 μA to 20 mA	0.3
1.24 to 5.3 (Adj.)	LM285BY	I	±1%	50	−40°C to +85°C	10 μA to 20 mA	0.3
1.24 to 5.3 (Adj.)	LM285	I	±2%	150	−40°C to +85°C	10 μA to 20 mA	0.3
1.24 to 5.3 (Adj.)	LM385BX	C	±1%	30	0°C to +70°C	13 μA to 20 mA	0.3
1.24 to 5.3 (Adj.)	LM385BY	C	±1%	50	0°C to +70°C	13 μA to 20 mA	0.3
1.24 to 5.3 (Adj.)	LM385	C	±2%	150	0°C to +70°C	13 μA to 20 mA	0.3
1.24 to 6.3 (Adj.)	†LM611M	M	±0.4%	20	−55°C to +125°C	16 μA to 10 mA	0.27
1.24 to 6.3 (Adj.)	LM611AI	I	±0.6%	20	−40°C to +85°C	16 μA to 10 mA	0.27
1.24 to 6.3 (Adj.)	LM611I	I	±0.6%	80	−40°C to +85°C	16 μA to 10 mA	0.27
1.24 to 6.3 (Adj.)	LM611C	C	±2.0%	150	0°C to +70°C	16 μA to 10 mA	0.27
1.24 to 6.3 (Adj.)	††LM613M	M	±0.4%	20	−55°C to +125°C	16 μA to 10 mA	0.2
1.24 to 6.3 (Adj.)	LM613AI	I	±0.6%	20	−40°C to +85°C	16 μA to 10 mA	0.2
1.24 to 6.3 (Adj.)	LM613I	I	±0.6%	80	−40°C to +85°C	16 μA to 10 mA	0.2
1.24 to 6.3 (Adj.)	LM613C	C	±2.0%	150	0°C to +70°C	16 μA to 10 mA	0.2
1.24 to 6.3 (Adj.)	‡LM614M	M	±0.4%	20	−55°C to +125°C	16 μA to 10 mA	0.2
1.24 to 6.3 (Adj.)	LM614AI	I	±0.6%	20	−40°C to +85°C	16 μA to 10 mA	0.2
1.24 to 6.3 (Adj.)	LM614I	I	±0.6%	80	−40°C to +85°C	16 μA to 10 mA	0.2
1.24 to 6.3 (Adj.)	LM614C	C	±2.0%	150	0°C to +70°C	16 μA to 10 mA	0.2
2.49	LM136A	M	±1%	72	−55°C to +125°C	400 μA to 10 mA	0.4
2.49	LM136	M	±2%	72	−55°C to +125°C	400 μA to 10 mA	0.4
2.49	LM236A	I	±1%	72	−25°C to +85°C	400 μA to 10 mA	0.4
2.49	LM236	I	±2%	72	−25°C to +85°C	400 μA to 10 mA	0.4
2.49	LM336	I	±4%	54	0°C to +70°C	400 μA to 10 mA	0.4
2.49	LM336B	C	±2%	54	0°C to +70°C	400 μA to 10 mA	0.4

Shunt Type (Continued)

Reverse Breakdown Voltage (V_R)	Device	Operating Temp. Range*	Voltage Tolerance Max, T_A = 25°C	Temperature Drift		Operating Current Range, I_R	Output Dynamic Impedance (Typ)
				ppm/°C (Max)	Over Range		
2.5	LM185BX-2.5	M	±1.5%	30	−55°C to +125°C	20 μA to 20 mA	1
2.5	LM185BY-2.5	M	±1.5%	50	−55°C to +125°C	20 μA to 20 mA	1
2.5	LM185B-2.5	M	±1.5%	150	−55°C to +125°C	20 μA to 20 mA	1
2.5	LM285BX-2.5	I	±1.5%	30	−40°C to +85°C	20 μA to 20 mA	1
2.5	LM285BY-2.5	I	±1.5%	50	−40°C to +85°C	20 μA to 20 mA	1
2.5	LM285-2.5	I	±1.5%	150	−40°C to +85°C	20 μA to 20 mA	1
2.5	LM385BX-2.5	C	±1.5%	30	0°C to +70°C	20 μA to 20 mA	1
2.5	LM385BY-2.5	C	±1.5%	50	0°C to +70°C	20 μA to 20 mA	1
2.5	LM385B-2.5	C	±1.5%	150	0°C to +70°C	20 μA to 20 mA	1
2.5	LM385-2.5	C	±3%	150	0°C to +70°C	20 μA to 20 mA	1
5.0	LM136A	M	±1%	72	−55°C to +125°C	400 μA to 10 mA	0.8
5.0	LM136	M	±2%	72	−55°C to +125°C	400 μA to 10 mA	0.8
5.0	LM236A	I	±1%	72	−25°C to +85°C	400 μA to 10 mA	0.8
5.0	LM236	I	±2%	72	−25°C to +85°C	400 μA to 10 mA	0.8
5.0	LM336B	C	±2%	54	0°C to +70°C	400 μA to 10 mA	0.8
5.0	LM336	C	±4%	54	0°C to +70°C	400 μA to 10 mA	0.8
6.9	LM129A	M	+3%, −2%	10	−55°C to +125°C	600 μA to 15 mA	0.6
6.9	LM129B	M	+3%, −2%	20	−55°C to +125°C	600 μA to 15 mA	0.6
6.9	LM129C	M	+3%, −2%	50	−55°C to +125°C	600 μA to 15 mA	0.6
6.9	LM329B	C	±5%	50	0°C to +70°C	600 μA to 15 mA	0.8
6.9	LM329C	C	±5%	20	0°C to +70°C	600 μA to 15 mA	0.8
6.9	LM329D	C	±5%	100	0°C to +70°C	600 μA to 15 mA	0.8
6.95	LM199A	M	±2%	0.5	−55°C to +125°C	500 μA to 10 mA	0.5
6.95	LM199A-20	M	Same as LM199A with 20 ppm guaranteed long term drift.				
6.95	LM199	M	±2%	1.0	−55°C to +125°C	500 μA to 10 mA	0.5
6.95	LM299A	I	±2%	0.5	−25°C to +85°C	500 μA to 10 mA	0.5
6.95	LM299A-20	I	Same as LM299A with 20 ppm guaranteed long term drift.				
6.95	LM299	I	±2%	1	−25°C to +85°C	500 μA to 10 mA	0.5
6.95	LM399A	C	±5%	1	0°C to +70°C	500 μA to 10 mA	0.5
6.95	LM399A-50	C	Same as LM399A with 50 ppm guaranteed long term drift.				
6.95	LM399	C	±5%	2	0°C to +70°C	500 μA to 10 mA	0.5
6.95	LM3999	C	±5%	5	0°C to +70°C	600 μA to 10 mA	0.6

*C (Commercial) = 0°C to 70°C, I (Industrial) = −25°C to +85°C for the LM236 and LM299, I = −40°C to +85°C for all others.
M (Military) = −55°C to +125°C

†LM611 has on-board Op Amp.

††LM613 has on-board Dual Op Amp and Dual Comparator.

‡LM614 has on-board Quad Op Amp.

Current References

Output Current Range	Device	Operating Temperature Range	Set Current Error			Operating Voltage Range	Set Current Temperature Dependence*
			2 μA to 10 μA	10 μA to 1 mA	1 mA to 5 mA		
2 μA to 10 mA	LM134	−55°C to +125°C	±8%	±3%	±5%	1V to 40V	0.96T to 0.104T
2 μA to 10 mA	LM134-3	−55°C to +125°C	N/A	±1%	N/A	1V to 40V	0.98T to 0.102T
2 μA to 10 mA	LM134-6	−55°C to +125°C	N/A	±2%	N/A	1V to 40V	0.97T to 0.103T
2 μA to 10 mA	LM234	−25°C to +100°C	±8%	±3%	±5	1V to 40V	0.96T to 0.104T
2 μA to 10 mA	LM234-3	−25°C to +100°C	N/A	±1%	N/A	1V to 40V	0.98T to 0.102T
2 μA to 10 mA	LM234-6	−25°C to +100°C	N/A	±2%	N/A	1V to 40V	0.97T to 0.103T
2 μA to 10 mA	LM334	0°C to +70°C	±12%	±6%	±8%	1V to 40V	0.96T to 0.104T

*Set current changes linearly with temperature at a rate of 0.33%/°C.

Series Type (Buffered Output)

Output Voltage	Device	Oper. Temp. Range*	Voltage Tolerance Max, T_A = 25°C	Temperature Drift		Load Reg. ppm/mA	Operating Current Range	Quiescent Current (mA)
				ppm/°C (Max)	Over Range			
0.2 (Adj)	†LM10	M	±2.5%	20 typ	−55°C to +125°C	100	0 mA to +1 mA	0.27
0.2 (Adj)	†LM10B	I	±2.5%	20 typ	−25°C to +85°C	100	0 mA to +1 mA	0.27
0.2 (Adj)	†LM10C	C	±5.0%	30 typ	0°C to +70°C	100	0 mA to +1 mA	0.30
2.5	LM368Y-2.5	C	±0.2%	20	0°C to +70°C	25	0 mA to +10 mA	0.55
2.5	LM368-2.5	C	±0.2%	30	0°C to +70°C	25	0 mA to +10 mA	0.55
5.0	LM168BY-5.0	M	±0.05%	10	−55°C to +125°C	10	−10 mA to +10 mA	0.35
5.0	LM268BY-5.0	I	±0.05%	15	−40°C to +85°C	10	−10 mA to +10 mA	0.35
5.0	LM368BY-5.0	C	±0.1%	20	0°C to +70°C	10	−10 mA to +10 mA	0.35
5.0	LM368-5.0	C	±0.1%	30	0°C to +70°C	10	−10 mA to +10 mA	0.35
10	LM169B	M	±0.05%	3	−55°C to +125°C	8	−10 mA to +10 mA	1.8
10	LH0070-2	M	±0.05%	8	−25°C to +25°C	60	0 to 5 mA	5
10	LM169	M	±0.05%	5	−55°C to +125°C	8	−10 mA to +10 mA	1.8
10	LH0070-0	M	±0.1%	40	−25°C to +25°C	60	0 mA to 5 mA	5
10	LH0070-1	M	±0.1%	20	−25°C to +25°C	60	0 mA to 5 mA	5
10	LM369C	C	±0.05%	10	0°C to +70°C	8	−10 mA to +10 mA	1.8
10	LM369	C	±0.05%	5	0°C to +70°C	8	−10 mA to +10 mA	1.8
10	LM369B	C	±0.05%	3	0°C to +70°C	8	−10 mA to +10 mA	1.8
10	LM368Y-10	C	±0.1%	20	0°C to +70°C	10	−10 mA to +10 mA	0.35
10	LM368-10	C	±0.1%	30	0°C to +70°C	10	−10 mA to +10 mA	0.35
10	LM369D	C	±0.1%	30	0°C to +70°C	8	−10 mA to +10 mA	2
10.24	LH0071-2	M	±0.05%	8	−40°C to +85°C	60	0 mA to 5 mA	5
10.24	LH0071-1	M	±0.1%	20	−40°C to +85°C	60	0 mA to 5 mA	5
10.24	LH0071-0	M	±0.1%	40	−25°C to +25°C	60	0 mA to 5 mA	5

*C (Commercial) = 0°C to 70°C, I (Industrial) = −40°C to +85°C, M (Military) = −55°C to +125°C

†Reference has on-board Op Amp.

Low Current Reference Diodes

Output Voltage	Device	Operating Temp. Range*	Voltage Tolerance Max, T_A = 25°C	Temperature Drift		Operating Current Range, I_R	Output Dynamic Impedance (Typ)
				ppm/°C (Max)	Over Range		
3.0	LM103-3.0	M	±10%	−1700	−55°C to +125°C	10 μA to 10 mA	25
3.3	LM103-3.3	M	±10%	−1500	−55°C to +125°C	10 μA to 10 mA	25
3.6	LM103-3.6	M	±10%	−1400	−55°C to +125°C	10 μA to 10 mA	25
3.9	LM103-3.9	M	±10%	−1300	−55°C to +125°C	10 μA to 10 mA	25

*M (Military) = −55°C to +125°C

7

"Reference Grade" Voltage Regulators*

Output Voltage	Device	Operating Temperature Range	Voltage Tolerance Max, $T_A = 25°C$	Output Variation Over Operating Range	Load Reg. ppm/mA	Line Reg. ppm/V	Output Current (Max)	Quiescent Current
Adjustable:	LP2951	−55°C to +150°C	±0.5%	±0.5%	100	42	100 mA	120 μA
1.235V to 30V	LP2951AC	−40°C to +125°C	±0.5%	±0.5%	100	42	100 mA	120 μA
	LP2951C	−40°C to +125°C	±1%	±1%	200	83	100 mA	120 μA
Programmable:	LH0075	−55°C to +125°C	±0.5%	±0.3% (Typ)	38	200	200 mA	8 mA
5V, 6V, 10V, 12V, 15V	LH0075C	0°C to +70°C	±1%	±0.14% (Typ)	76	400	200 mA	10 mA
Programmable								
−5V, −6V, −10V	LH0076	−55°C to +125°C	±0.5%	±0.3% (Typ)	38	200	200 mA	15 mA
−10V, −15V	LH0076C	0°C to +70°C	±1%	±0.14% (Typ)	38	400	200 mA	15 mA
5V	LP2950AC	−40°C to +125°C	±0.5%	±0.5%	100	42	100 mA	120 μA
5V	LP2950C	−40°C to +125°C	±1%	±1%	200	83	100 mA	120 μA

*For more information on these circuits, refer to the Continuous Voltage Regulators section of the General Purpose Linear Devices Databook.

National Semiconductor

LH0070 Series Precision BCD Buffered Reference
LH0071 Series Precision Binary Buffered Reference

General Description

The LH0070 and LH0071 are precision, three terminal, voltage references consisting of a temperature compensated zener diode driven by a current regulator and a buffer amplifier. The devices provide an accurate reference that is virtually independent of input voltage, load current, temperature and time. The LH0070 has a 10.000V nominal output to provide equal step sizes in BCD applications. The LH0071 has a 10.240V nominal output to provide equal step sizes in binary applications.

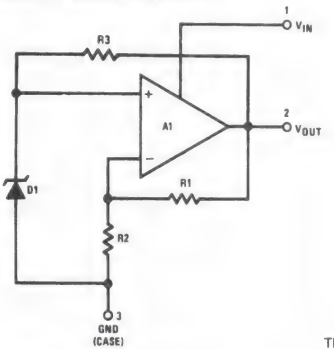
The output voltage is established by trimming ultra-stable, low temperature drift, thin film resistors under actual operating circuit conditions. The devices are shortcircuit proof in both the current sourcing and sinking directions.

The LH0070 and LH0071 series combine excellent long term stability, ease of application, and low cost, making them ideal choices as reference voltages in precision D to A and A to D systems.

Features

- Accuracy output voltage
  - LH0070 10V ± 0.02%
  - LH0071 10.24V ± 0.02%
- Single supply operation 11.4V to 40V
- Low output impedance 0.2Ω
- Excellent line regulation 0.1 mV/V
- Low zener noise 20 μVp-p
- 3-lead TO-5 (pin compatible with the LM109)
- Short circuit proof
- Low standby current 3 mA

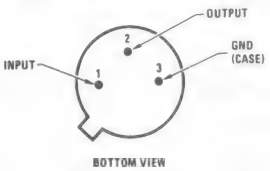
Equivalent Schematic



TL/H/5550–1

Connection Diagram

TO-5 Metal Can Package

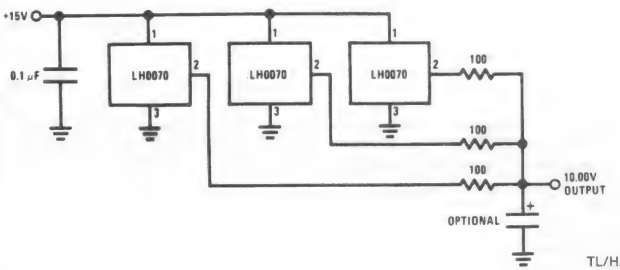


TL/H/5550–7

Order Number LH0070-0H, LH0071-0H, LH0070-1H, LH0071-1H, LH0070-2H or LH0071-2H
See NS Package Number H03B

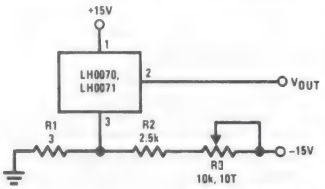
Typical Applications

Statistical Voltage Standard



TL/H/5550–8

***Output Voltage Fine Adjustment**



TL/H/5550–9

***Note:** The output of the LH0070 and LH0071 may be adjusted to a precise voltage by using the above circuit since the supply current of the devices is relatively small and constant with temperature and input voltage. For the circuit shown, supply sensitivities are degraded slightly to 0.01%/V change in V_{OUT} for changes in V_{IN} and V^-.

An additional temperature drift of 0.0001%/°C is added due to the variation of supply current with temperature of the LH0070 and LH0071. Sensitivity to the value of R1, R2 and R3 is less than 0.001%/%.

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

(Note 4)

Supply Voltage	40V
Power Dissipation (See Curve)	600 mW
Short Circuit Duration	Continuous
Output Current	± 20 mA
Operating Temperature Range	−55°C to +125°C
Storage Temperature Range	−65°C to ±150°C
Lead Temp. (Soldering, 10 seconds)	300°C

Electrical Characteristics (Note 1)

Parameter	Conditions	Min	Typ	Max	Units
Output Voltage	$T_A = 25°C$				
LH0070			10.000		V
LH0071			10.24		V
Output Accuracy	$T_A = 25°C$				
−0, −1			±0.03	±0.1	%
−2			±0.02	±0.05	%
Output Accuracy	$T_A = -55°C, 125°C$				
−0, −1				±0.3	%
−2				±0.2	%
Output Voltage Change With Temperature	(Note 2)				
−0				±0.2	%
−1			±0.02	±0.1	%
−2			±0.01	±0.04	%
Line Regulation	$13V \le V_{IN} \le 33V, T_C = 25°C$				
−0, −1			0.02	0.1	%
−2			0.01	0.03	%
Input Voltage Range	$R_L = 50\ k\Omega$	11.4		40	V
Load Regulation	$0\ mA \le I_{OUT} \le 5\ mA$		0.01	0.03	%
Quiescent Current	$13V \le V_{IN} \le 33V, I_{OUT} = 0\ mA$	1	3	5	mA
Change In Quiescent Current	$\Delta V_{IN} = 20V$ From 23V To 33V		0.75	1.5	mA
Output Noise Voltage	BW = 0.1 Hz To 10 Hz, $T_A = 25°C$		20		μVp-p
Ripple Rejection	f = 120 Hz		0.01		%/Vp-p
Output Resistance			0.2	0.6	Ω
Long Term Stability	$T_A = 25°C$ (Note 3)				
−0, −1				±0.2	%/yr.
−2				±0.05	%/yr.
Thermal Resistance	$T_j = 150°C$				
θ_{ja} (Junction to Ambient)			200		°C/W
θ_{jc} (Junction to Case)			100		°C/W

Note 1: Unless otherwise specified, these specifications apply for $V_{IN} = 15.0V$, $R_L = 10\ k\Omega$, and over the temperature range of $-55°C \le T_A \le +125°C$.

Note 2: This specification is the difference in output voltage measured at $T_A = 85°C$ and $T_A = 25°C$ or $T_A = 25°C$ and $T_A = -25°C$ with readings taken after test chamber and device-under-test stabilization at temperature using a suitable precision voltmeter.

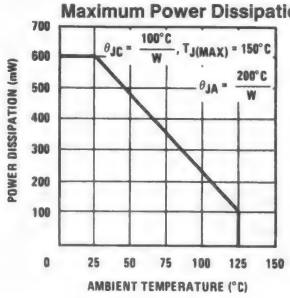
Note 3: This parameter is guaranteed by design and not tested.

Note 4: Refer to the following RETS drawings for military specifications:

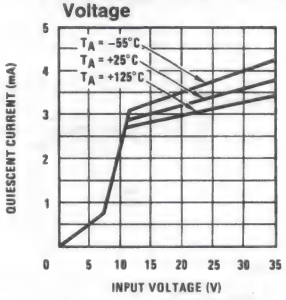
RETS0070-0H for LH0070-0H	RETS0071-0H for LH0071-0H
RETS0070-1H for LH0070-1H	RETS0071-1H for LH0071-1H
RETS0070-2H for LH0070-2H	RETS0071-2H for LH0071-2H

Typical Performance Characteristics

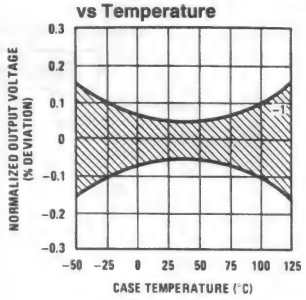
Maximum Power Dissipation



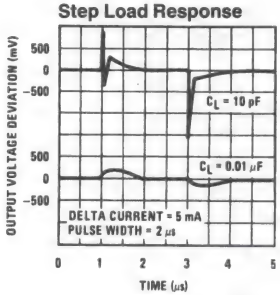
Quiescent Current vs Input Voltage



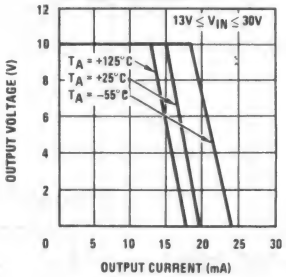
Normalized Output Voltage vs Temperature



Step Load Response

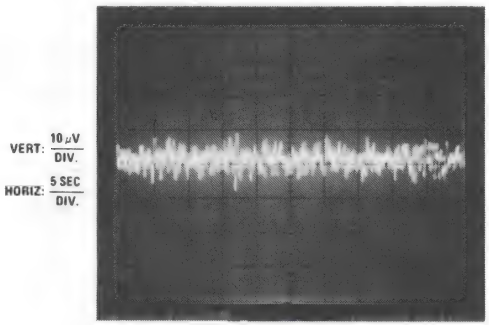


Output Short Circuit Characteristics



TL/H/5550–2

Noise Voltage

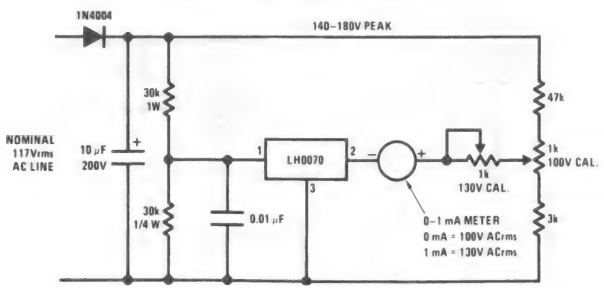


BW = 0.1 Hz TO 10 Hz

TL/H/5550–6

Typical Applications (Continued)

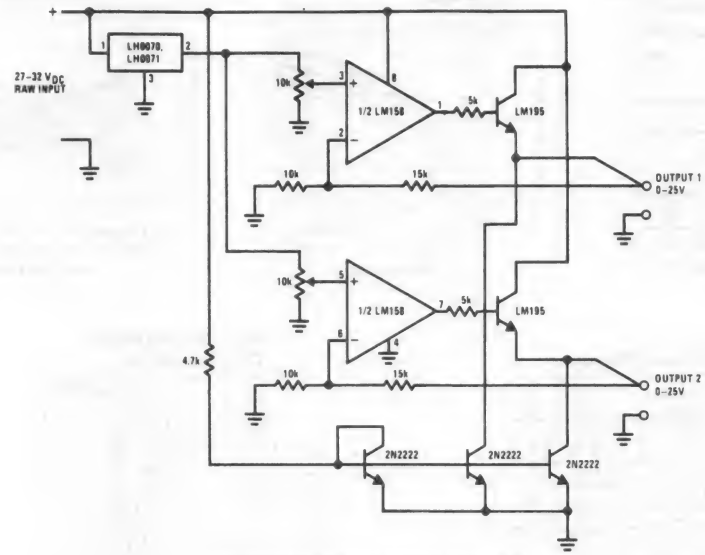
Expanded Scale AC Voltmeter



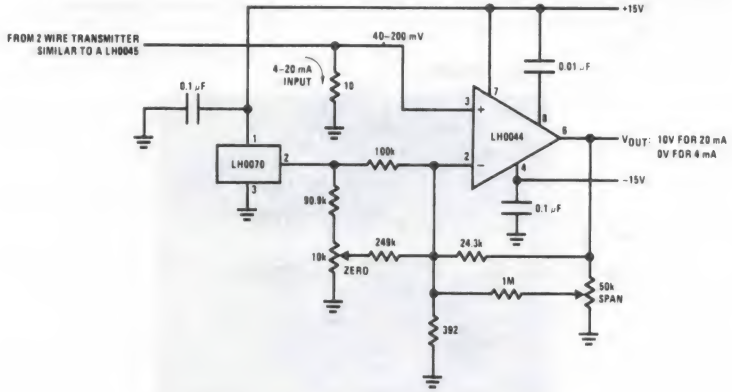
TL/H/5550–4

Typical Applications (Continued)

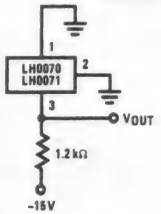
Dual Output Bench Power Supply



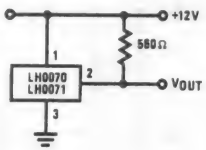
Precision Process Control Interface



Negative 10V Reference



Boosted Reference For Low Input Voltages



TL/H/5550–5

National Semiconductor

LH7070 Series Precision BCD Buffered Reference
LH7071 Series Precision Binary Buffered Reference

General Description

The LH7070 and LH7071 are precision, three terminal, voltage references consisting of a temperature compensated zener diode driven by a current regulator and a buffer amplifier. The devices provide an accurate reference that is virtually independent of input voltage, load current, temperature and time. The LH7070 has a 10.000V nominal output to provide equal step sizes in BCD applications. The LH7071 has a 10.240V nominal output to provide equal step sizes in binary applications.

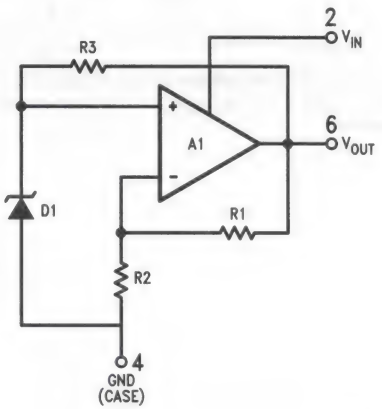
The output voltage is established by trimming ultrastable, low temperature drift, thin film resistors under actual operating circuit conditions. The devices are short-circuit proof in both the current sourcing and sinking directions.

The LH7070 and LH7071 series combine excellent long term stability, ease of application, and low cost, making them ideal choices as reference voltages in precision D to A and A to D systems.

Features

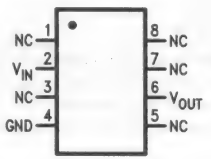
- Accurate output voltage
  - — LH7070 10V ±0.03%
  - — LH7071 10.24V ±0.03%
- Single supply operation 11.4V to 40V
- Low output impedance 0.2Ω
- Excellent line regulation 0.2 mV/V
- Low zener noise 20 μVp-p
- Short circuit proof
- Low standby current 3 mA

Equivalent Schematic



TL/K/10032–1

Connection Diagram



TL/K/10032–2

Top View

Order Number LH7070CN or LH7071CN
See NS Package Number N08E

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	40V
Power Dissipation (See Curve)	800 mW
Short Circuit Duration	Continuous
Output Current	±20 mA
Operating Temperature Range	−25°C to +85°C
Storage Temperature Range	−65°C to +150°C
Lead Temperature (Soldering, 10 Sec.)	300°C

Electrical Characteristics (Note 1)

Parameter	Conditions	Min	Typ	Max	Units
Output Voltage	$T_A = 25°C$				
	LH7070		10.000		V
	LH7071		10.240		V
Output Accuracy	$T_A = 25°C$ LH7070, LH7071		±0.03	±0.1	%
Output Accuracy	$T_A = -25°C$ to $+85°C$ (Note 3)			±0.3	%
Output Voltage Change with Temperature	(Notes 2, 3)			±0.14	%
Line Regulation	$13V \le V_{IN} \le 33V$, $T_A = 25°C$		0.02	0.1	%
Input Voltage Range	$R_L = 50\ k\Omega$	11.4		40	V
Load Regulation	$0\ mA \le I_{OUT} \le 5\ mA$		0.01	0.03	%
Quiescent Current	$13V \le V_{IN} \le 33V$, $I_{OUT} = 0\ mA$	1	2	3	mA
Change in Quiescent Current	$\Delta V_{IN} = 20V$ from 13V to 33V		0.75	1.5	mA
Output Noise Voltage	BW = 0.1 Hz to 10 Hz, $T_A = 25°C$		20		μV_{p-p}
Ripple Rejection	f = 120 Hz		0.01		$\%/V_{p-p}$
Output Resistance			0.2	0.6	Ω
Long Term Stability	$T_A = 25°C$ (Note 3)			±0.2	%/yr.

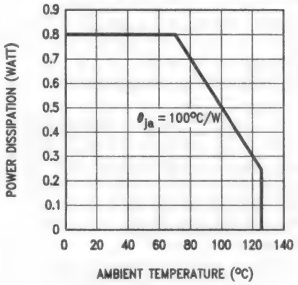
Note 1: Unless otherwise specified, these specifications apply for $V_{IN} = 15.0V$, $R_L = 10\ k\Omega$, and over the temperature range of $-25°C \le T_A \le +85°C$.

Note 2: This specification is the difference in output voltage measured at $T_A = 85°C$ and $T_A = 25°C$ or $T_A = 25°C$ and $T_A = -25°C$ with readings taken after test chamber and device-under-test stabilization at temperature using a suitable precision voltmeter.

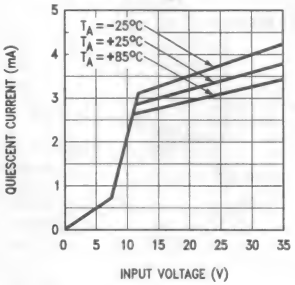
Note 3: This parameter is guaranteed by design and not tested.

Typical Performance Characteristics

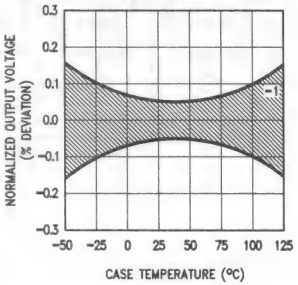
Maximum Power Dissipation



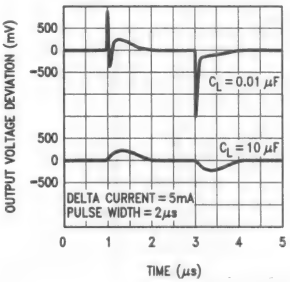
Quiescent Current vs Input Voltage



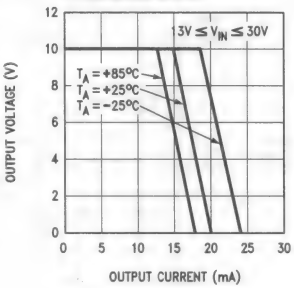
Normalized Output Voltage vs Temperature



Load Transient Response

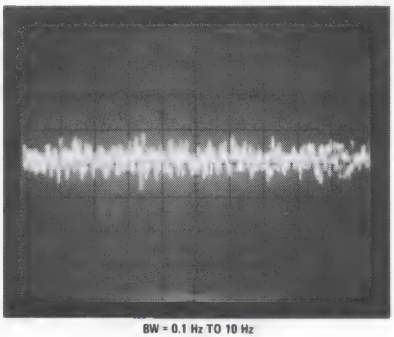


Output Short Circuit Characteristics



TL/K/10032–5

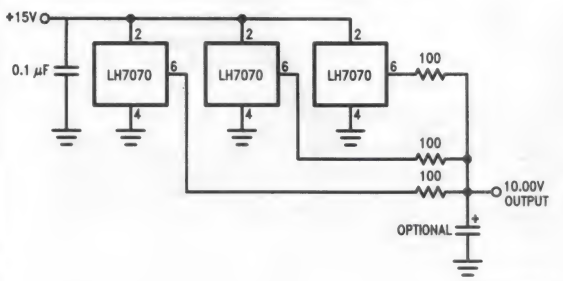
Noise Voltage



TL/K/10032–6

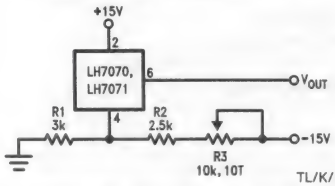
Typical Applications

Statistical Voltage Standard



TL/K/10032–3

*Output Voltage Fine Adjustment

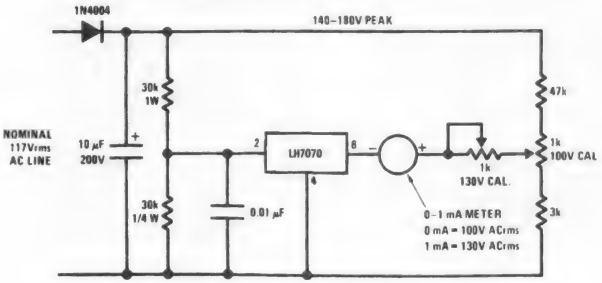


TL/K/10032–4

***Note:** The output of the LH7070 and LH7071 may be adjusted to a precise voltage by using the above circuit since the supply current of the device is relatively small and constant with temperature and input voltage. For the circuit shown, supply sensitivities are degraded slightly to 0.01%/V change in V_{OUT} for changes in V_{IN} and V^-.

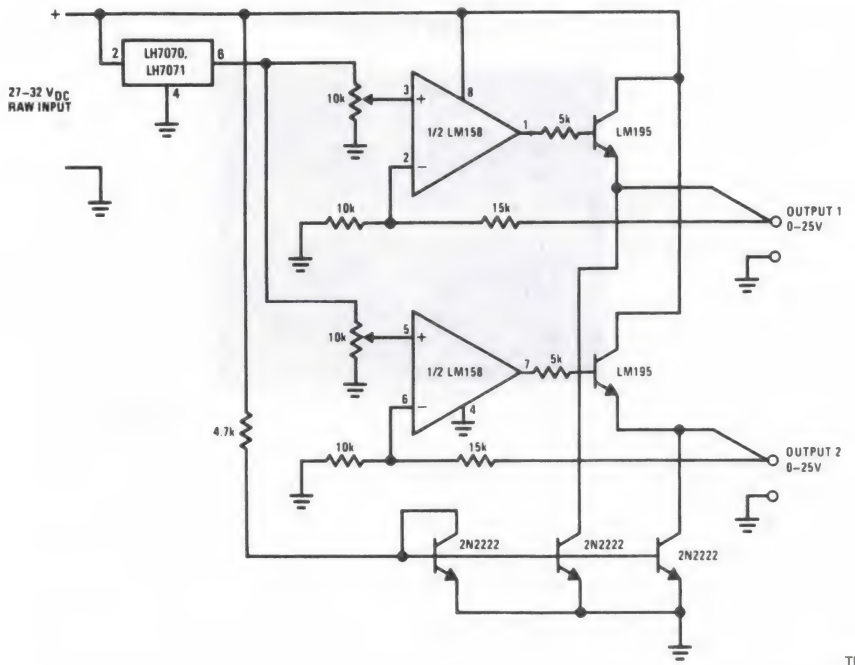
An additional temperature drift of 0.0001%/°C is added due to the variation of supply current with temperature of the LH7070 and LH7071. Sensitivity to the value of R1, R2 and R3 is less than 0.001%/%.

Expanded Scale AC Voltmeter



TL/K/10032–7

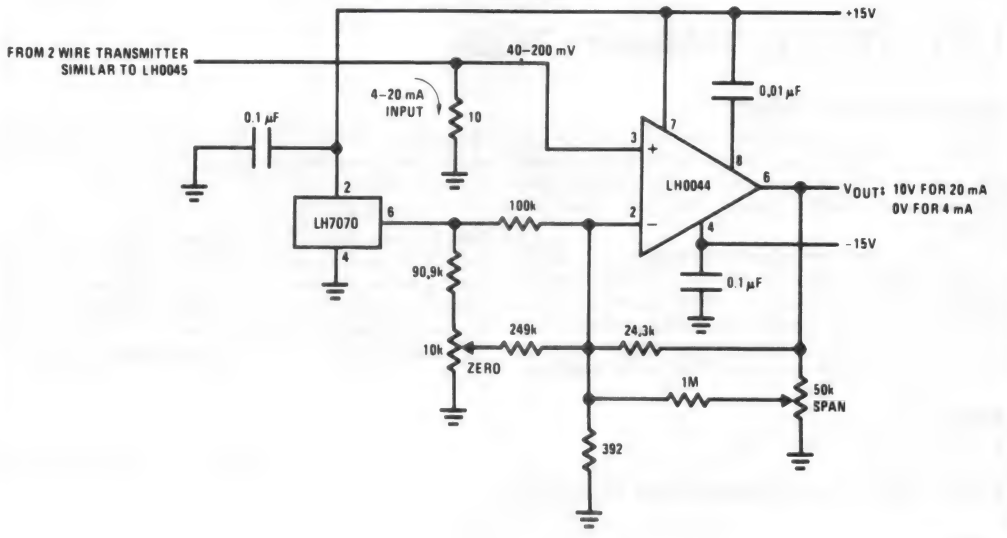
Dual Output Bench Power Supply



TL/K/10032–8

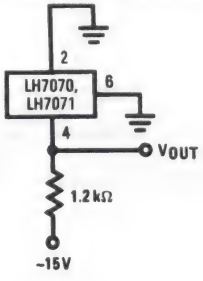
Typical Applications (Continued)

Precision Process Control Interface



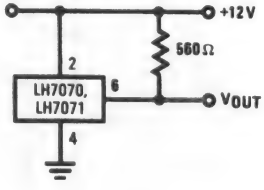
TL/K/10032–9

Negative 10V Reference



TL/K/10032–10

Current Boost for Low Input Voltages



TL/K/10032–11

National Semiconductor

LM113/LM313 Reference Diode

General Description

The LM113/LM313 are temperature compensated, low voltage reference diodes. They feature extremely-tight regulation over a wide range of operating currents in addition to an unusually-low breakdown voltage and good temperature stability.

The diodes are synthesized using transistors and resistors in a monolithic integrated circuit. As such, they have the same low noise and long term stability as modern IC op amps. Further, output voltage of the reference depends only on highly-predictable properties of components in the IC; so they can be manufactured and supplied to tight tolerances.

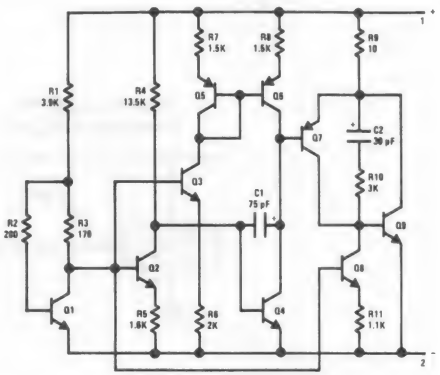
- Dynamic impedance of 0.3Ω from 500 μA to 20 mA
- Temperature stability typically 1% over −55°C to 125°C range (LM113), 0°C to 70°C (LM313)
- Tight tolerance: ±5%, ±2% or ±1%

The characteristics of this reference recommend it for use in bias-regulation circuitry, in low-voltage power supplies or in battery powered equipment. The fact that the breakdown voltage is equal to a physical property of silicon—the energy-band gap voltage—makes it useful for many temperature-compensation and temperature-measurement functions.

Features

- Low breakdown voltage: 1.220V

Schematic and Connection Diagrams



Metal Can Package

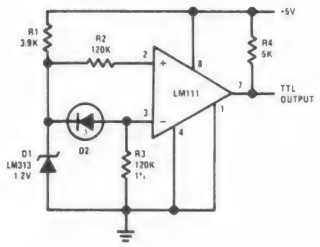
Note: Pin 2 connected to case.
TOP VIEW

Order Number LM113H or LM113-1H or LM113-2H or LM313H
See NS Package Number H02A

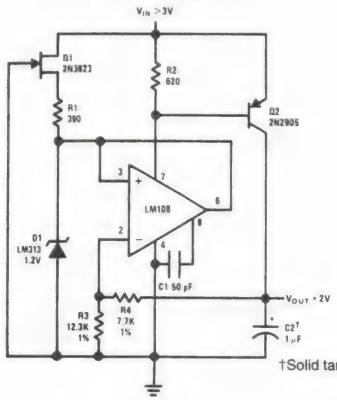
TL/H/5713–1

Typical Applications

Level Detector for Photodiode



Low Voltage Regulator



†Solid tantalum.

TL/H/5713–2

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications. (Note 3)

Power Dissipation (Note 1)	100 mW
Reverse Current	50 mA
Forward Current	50 mA
Storage Temperature Range	−65°C to +150°C
Lead Temperature (Soldering, 10 seconds)	300°C
Operating Temperature Range	
LM113	−55°C to +125°C
LM313	0°C to +70°C

Electrical Characteristics (Note 2)

Parameter	Conditions	Min	Typ	Max	Units
Reverse Breakdown Voltage					
LM113/LM313	I_R = 1 mA	1.160	1.220	1.280	V
LM113-1		1.210	1.22	1.232	V
LM113-2		1.195	1.22	1.245	V
Reverse Breakdown Voltage Change	0.5 mA ≤ I_R ≤ 20 mA		6.0	15	mV
Reverse Dynamic Impedance	I_R = 1 mA		0.2	1.0	Ω
	I_R = 10 mA		0.25	0.8	Ω
Forward Voltage Drop	I_F = 1.0 mA		0.67	1.0	V
RMS Noise Voltage	10 Hz ≤ f ≤ 10 kHz, I_R = 1 mA		5		μV
Reverse Breakdown Voltage Change with Current	0.5 mA ≤ I_R ≤ 10 mA, T_{MIN} ≤ T_A ≤ T_{MAX}			15	mV
Breakdown Voltage Temperature Coefficient	1.0 mA ≤ I_R ≤ 10 mA, T_{MIN} ≤ T_A ≤ T_{MAX}		0.01		%/°C

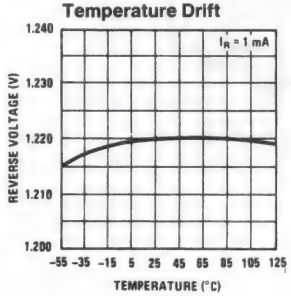
Note 1: For operating at elevated temperatures, the device must be derated based on a 150°C maximum junction and a thermal resistance of 80°C/W junction to case or 440°C/W junction to ambient.

Note 2: These specifications apply for T_A = 25°C, unless stated otherwise. At high currents, breakdown voltage should be measured with lead lengths less than ¼ inch. Kelvin contact sockets are also recommended. The diode should not be operated with shunt capacitances between 200 pF and 0.1 μF, unless isolated by at least a 100Ω resistor, as it may oscillate at some currents.

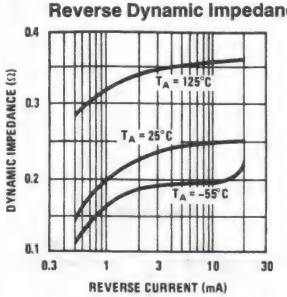
Note 3: Refer to the following RETS drawings for military specifications: RETS113-1X for LM113-1, RETS113-2X for LM113-2 or RETS113X for LM113.

Typical Performance Characteristics

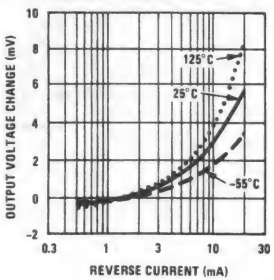
Temperature Drift



Reverse Dynamic Impedance



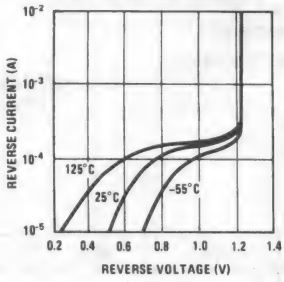
Reverse Characteristics



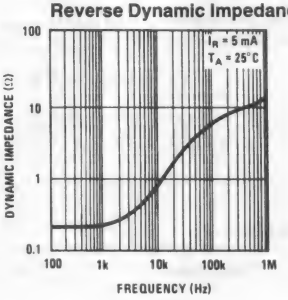
TL/H/5713–3

Typical Performance Characteristics (Continued)

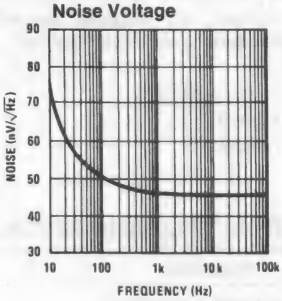
Reverse Characteristics



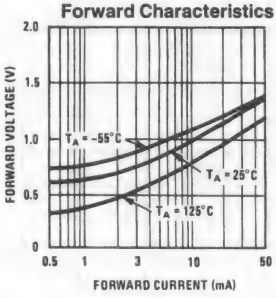
Reverse Dynamic Impedance



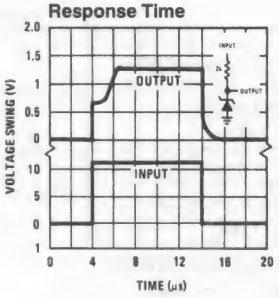
Noise Voltage



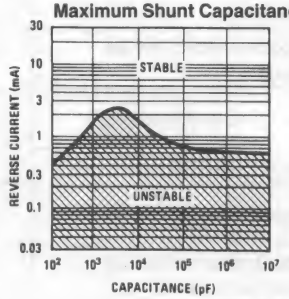
Forward Characteristics



Response Time



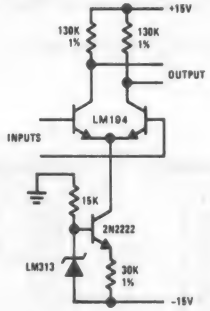
Maximum Shunt Capacitance



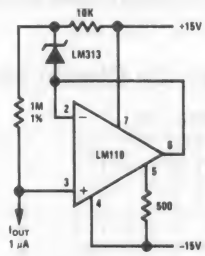
TL/H/5713–4

Typical Applications (Continued)

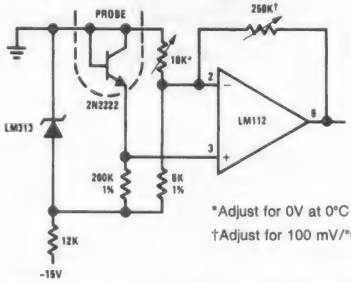
Amplifier Biasing for Constant Gain with Temperature



Constant Current Source



Thermometer



TL/H/5713–5

National Semiconductor

LM129/LM329 Precision Reference

General Description

The LM129 and LM329 family are precision multi-current temperature-compensated 6.9V zener references with dynamic impedances a factor of 10 to 100 less than discrete diodes. Constructed in a single silicon chip, the LM129 uses active circuitry to buffer the internal zener allowing the device to operate over a 0.5 mA to 15 mA range with virtually no change in performance. The LM129 and LM329 are available with selected temperature coefficients of 0.001, 0.002, 0.005 and 0.01%/°C. These new references also have excellent long term stability and low noise.

A new subsurface breakdown zener used in the LM129 gives lower noise and better long-term stability than conventional IC zeners. Further the zener and temperature compensating transistor are made by a planar process so they are immune to problems that plague ordinary zeners. For example, there is virtually no voltage shift in zener voltage due to temperature cycling and the device is insensitive to stress on the leads.

The LM129 can be used in place of conventional zeners with improved performance. The low dynamic impedance simplifies biasing and the wide operating current allows the replacement of many zener types.

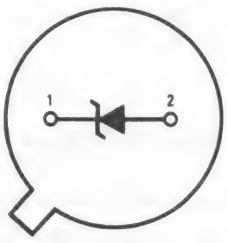
The LM129 is packaged in a 2-lead TO-46 package and is rated for operation over a $-55°C$ to $+125°C$ temperature range. The LM329 for operation over 0°C to 70°C is available in both a hermetic TO-46 package and a TO-92 epoxy package.

Features

- 0.6 mA to 15 mA operating current
- 0.6Ω dynamic impedance at any current
- Available with temperature coefficients of 0.001%/°C
- 7μV wideband noise
- 5% initial tolerance
- 0.002% long term stability
- Low cost
- Subsurface zener

Connection Diagrams

Metal Can Package



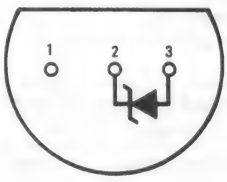
TL/H/5714–6

Bottom View

Pin 2 is electrically connected to case

Order Number LM129AH, LM129BH, LM129CH, LM329AH, LM329BH, LM329CH or LM329DH
See NS Package H02A

Plastic Package



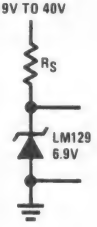
TL/H/5714–4

Bottom View

Order Number LM329BZ, LM329CZ or LM329DZ
See NS Package Z03A

Typical Applications

Simple Reference



TL/H/5714–1

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications. (Note 2)

Reverse Breakdown Current	30 mA
Forward Current	2 mA
Operating Temperature Range	
LM129	−55°C to +125°C
LM329	0°C to +70°C
Storage Temperature Range	−55°C to +150°C
Soldering Information	
TO-92 package: 10 sec.	260°C
TO-46 package: 10 sec.	300°C

Electrical Characteristics (Note 1)

Parameter	Conditions	LM129A, B, C			LM329A, B, C, D			Units
		Min	Typ	Max	Min	Typ	Max	
Reverse Breakdown Voltage	$T_A = 25°C$, $0.6\ mA \le I_R \le 15\ mA$	6.7	6.9	7.2	6.6	6.9	7.25	V
Reverse Breakdown Change with Current (Note 3)	$T_A = 25°C$, $0.6\ mA \le I_R \le 15\ mA$		9	14		9	20	mV
Reverse Dynamic Impedance (Note 3)	$T_A = 25°C$, $I_R = 1\ mA$		0.6	1		0.8	2	Ω
RMS Noise	$T_A = 25°C$, $10\ Hz \le F \le 10\ kHz$		7	20		7	100	μV
Long Term Stability (1000 hours)	$T_A = 45°C \pm 0.1°C$, $I_R = 1\ mA \pm 0.3\%$		20			20		ppm
Temperature Coefficient	$I_R = 1\ mA$							
LM129A, LM329A			6	10		6	10	ppm/°C
LM129B, LM329B			15	20		15	20	ppm/°C
LM129C, LM329C			30	50		30	50	ppm/°C
LM329D						50	100	ppm/°C
Change In Reverse Breakdown Temperature Coefficient	$1\ mA \le I_R \le 15\ mA$		1			1		ppm/°C
Reverse Breakdown Change with Current	$1\ mA \le I_R \le 15\ mA$		12			12		mV
Reverse Dynamic Impedance	$1\ mA \le I_R \le 15\ mA$		0.8			1		Ω

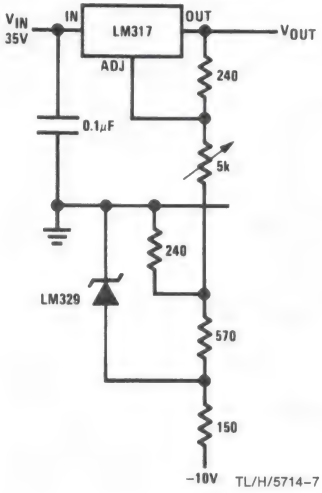
Note 1: These specifications apply for $-55°C \le T_A \le +125°C$ for the LM129 and $0°C \le T_A \le +70°C$ for the LM329 unless otherwise specified. The maximum junction temperature for an LM129 is 150°C and LM329 is 100°C. For operating at elevated temperature, devices in TO-46 package must be derated based on a thermal resistance of 440°C/W junction to ambient or 80°C/W junction to case. For the TO-92 package, the derating is based on 180°C/W junction to ambient with 0.4" leads from a PC board and 160°C/W junction to ambient with 0.125" lead length to a PC board.

Note 2: Refer to RETS129H for LM129 family military specifications.

Note 3: These changes are tested on a pulsed basis with a low duty-cycle. For changes versus temperature, compute in terms of tempco.

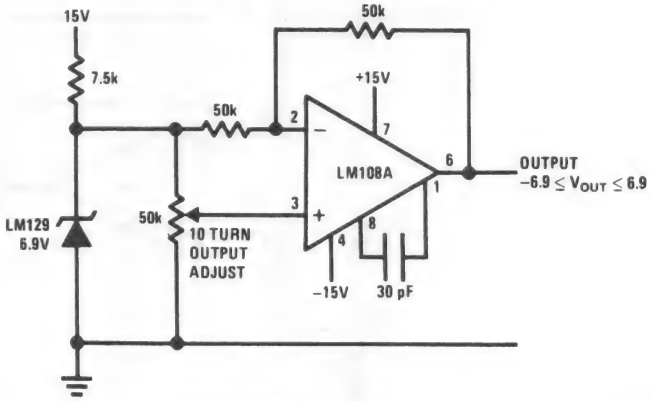
Typical Applications (Continued)

Low Cost 0–25V Regulator



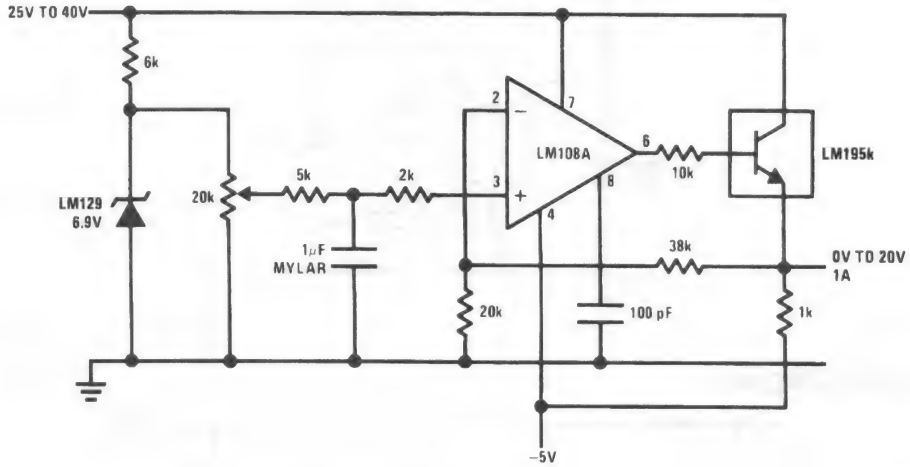
TL/H/5714–7

Adjustable Bipolar Output Reference



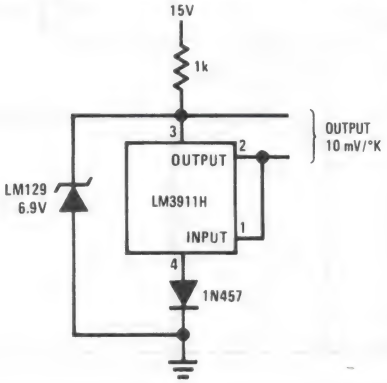
TL/H/5714–8

0V to 20V Power Reference



TL/H/5714–9

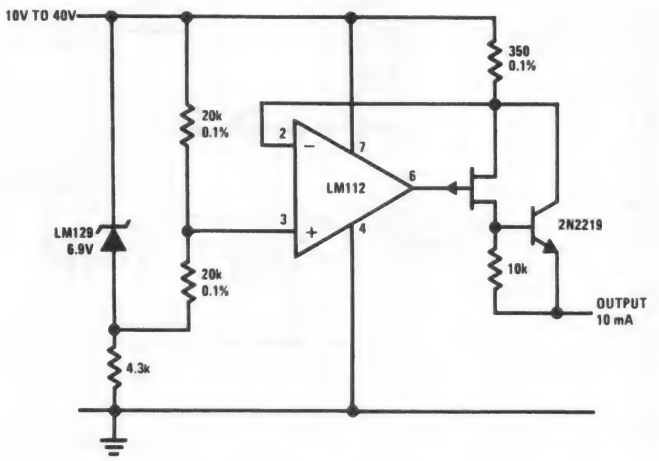
External Reference for Temperature Transducer



TL/H/5714–2

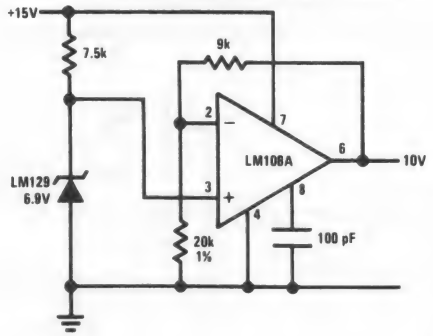
Typical Applications (Continued)

Positive Current Source



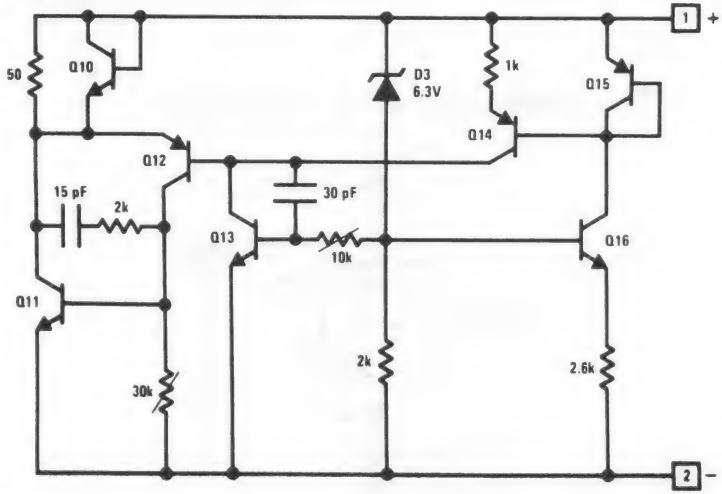
TL/H/5714–11

Buffered Reference with Single Supply



TL/H/5714–3

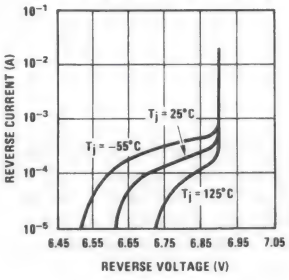
Schematic Diagram



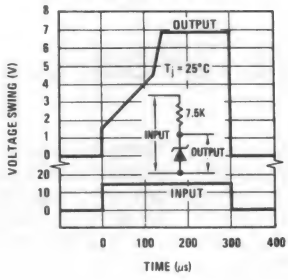
TL/H/5714–10

Typical Performance Characteristics

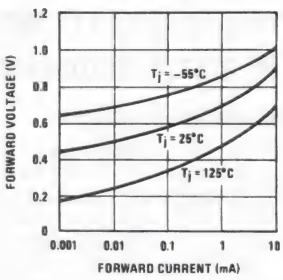
Reverse Characteristics



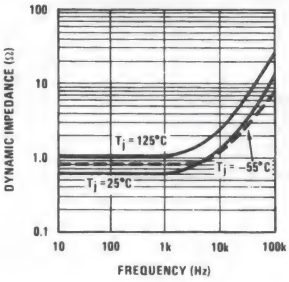
Response Time



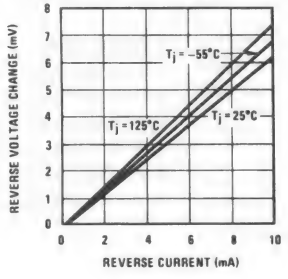
Forward Characteristics



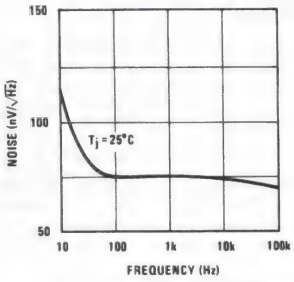
Dynamic Impedance



Reverse Voltage Change

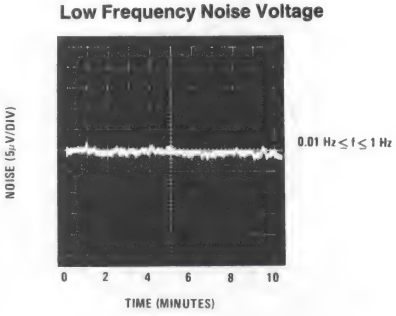


Zener Noise Voltage



TL/H/5714–12

Low Frequency Noise Voltage



TL/H/5714–5

7

National Semiconductor

LM134/LM234/LM334
3-Terminal Adjustable Current Sources

General Description

The LM134/LM234/LM334 are 3-terminal adjustable current sources featuring 10,000:1 range in operating current, excellent current regulation and a wide dynamic voltage range of 1V to 40V. Current is established with one external resistor and no other parts are required. Initial current accuracy is ±3%. The LM134/LM234/LM334 are true floating current sources with no separate power supply connections. In addition, reverse applied voltages of up to 20V will draw only a few dozen microamperes of current, allowing the devices to act as both a rectifier and current source in AC applications.

The sense voltage used to establish operating current in the LM134 is 64 mV at 25°C and is directly proportional to absolute temperature (°K). The simplest one external resistor connection, then, generates a current with ≈ +0.33%/°C temperature dependence. Zero drift operation can be obtained by adding one extra resistor and a diode.

Applications for the new current sources include bias networks, surge protection, low power reference, ramp generation, LED driver, and temperature sensing. The LM134-3/LM234-3 and LM134-6/LM234-6 are specified as true temperature sensors with guaranteed initial accuracy of ±3°C and ±6°C, respectively. These devices are ideal in remote sense applications because series resistance in long wire runs does not affect accuracy. In addition, only 2 wires are required.

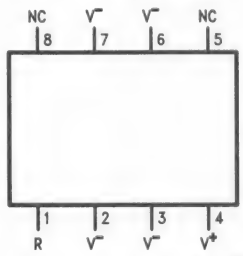
The LM134 is guaranteed over a temperature range of −55°C to +125°C, the LM234 from −25°C to +100°C and the LM334 from 0°C to +70°C. These devices are available in TO-46 hermetic, TO-92 and SO-8 plastic packages.

Features

- Operates from 1V to 40V
- 0.02%/V current regulation
- Programmable from 1 μA to 10 mA
- True 2-terminal operation
- Available as fully specified temperature sensor
- ±3% initial accuracy

Connection Diagrams

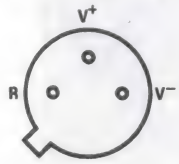
SO-8
Surface Mount Package



TL/H/5697–24

Order Number LM334M
See NS Package Number M08A

TO-46
Metal Can Package



TL/H/5697–12

Bottom View

Pin 3 is electrically connected to case.

Order Number LM134H, LM134H-3, LM134H-6, LM234H, LM234H-3, LM234H-6, or LM334H
See NS Package Number H03H

TO-92
Plastic Package

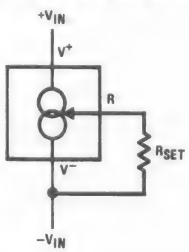
TL/H/5697–10

Bottom View

Order Number LM334Z, LM234Z-3 or LM234Z-6
See NS Package Number Z03A

Typical Application

Basic 2-Terminal Current Source



TL/H/5697–1

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

V^+ to V^- Forward Voltage	
LM134/LM234	40V
LM334/LM134-3/LM134-6/LM234-3/LM234-6	30V
V^+ to V^- Reverse Voltage	20V
R Pin to V^- Voltage	5V
Set Current	10 mA
Power Dissipation	400 mW

Operating Temperature Range (Note 4)	
LM134/LM134-3/LM134-6	−55°C to +125°C
LM234/LM234-3/LM234-6	−25°C to +100°C
LM334	0°C to +70°C
Soldering Information	
TO-92 Package (10 sec.)	260°C
TO-46 Package (10 sec.)	300°C
SO Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (Appendix D) for other methods of soldering surface mount devices.

Electrical Characteristics (Note 1)

Parameter	Conditions	LM134/LM234			LM334			Units
		Min	Typ	Max	Min	Typ	Max	
Set Current Error, $V^+ = 2.5V$, (Note 2)	$10\ \mu A \le I_{SET} \le 1$ mA			3			6	%
	1 mA $< I_{SET} \le 5$ mA			5			8	%
	$2\ \mu A \le I_{SET} < 10\ \mu A$			8			12	%
Ratio of Set Current to V^- Current	$100\ \mu A \le I_{SET} \le 1$ mA	14	18	23	14	18	26	
	1 mA $\le I_{SET} \le 5$ mA		14			14		
	$2\ \mu A \le I_{SET} \le 100\ \mu A$		18	23		18	26	
Minimum Operating Voltage	$2\ \mu A \le I_{SET} \le 100\ \mu A$		0.8			0.8		V
	$100\ \mu A < I_{SET} \le 1$ mA		0.9			0.9		V
	1 mA $< I_{SET} \le 5$ mA		1.0			1.0		V
Average Change in Set Current with Input Voltage	$2\ \mu A \le I_{SET} \le 1$ mA							
	$1.5V \le V^+ \le 5V$		0.02	0.05		0.02	0.1	%/V
	$5V \le V^+ \le 40V$		0.01	0.03		0.01	0.05	%/V
	1 mA $< I_{SET} \le 5$ mA							
	$1.5V \le V \le 5V$		0.03			0.03		%/V
	$5V \le V \le 40V$		0.02			0.02		%/V
Temperature Dependence of Set Current (Note 3)	$25\ \mu A \le I_{SET} \le 1$ mA	0.96T	T	1.04T	0.96T	T	1.04T	
Effective Shunt Capacitance			15			15		pF

Note 1: Unless otherwise specified, tests are performed at $T_j = 25°C$ with pulse testing so that junction temperature does not change during test.

Note 2: Set current is the current flowing into the V^+ pin. It is determined by the following formula: $I_{SET} = 67.7\ mV / R_{SET}$ (@ 25°C). Set current error is expressed as a percent deviation from this amount. I_{SET} increases at 0.336%/°C @ $T_j = 25°C$.

Note 3: I_{SET} is directly proportional to absolute temperature (°K). I_{SET} at any temperature can be calculated from: $I_{SET} = I_o\ (T/T_o)$ where I_o is I_{SET} measured at T_o (°K).

Note 4: For elevated temperature operation, T_j max is:

LM134	150°C
LM234	125°C
LM334	100°C

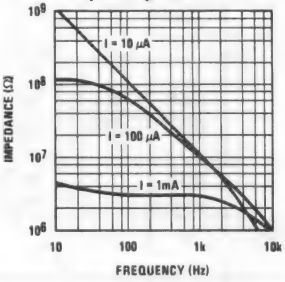
Thermal Resistance	TO-92	TO-46	SO-8
θ_{ja} (Junction to Ambient)	180°C/W (0.4" leads) 160°C/W (0.125" leads)	440°C/W	165°C/W
θ_{jc} (Junction to Case)	N/A	32°C/W	N/A

Electrical Characteristics (Note 1) (Continued)

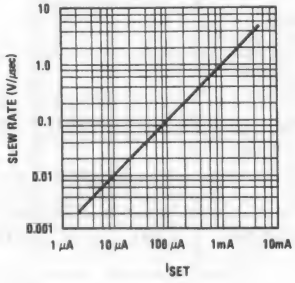
Parameter	Conditions	LM134-3, LM234-3			LM134-6, LM234-6			Units
		Min	Typ	Max	Min	Typ	Max	
Set Current Error, $V^+ = 2.5V$, (Note 2)	$100\ \mu A \leq I_{SET} \leq 1$ mA $T_j = 25°$			±1			±2	%
Equivalent Temperature Error				±3			±6	°C
Ratio of Set Current to V^- Current	$100\ \mu A \leq I_{SET} \leq 1$ mA	14	18	26	14	18	26	
Minimum Operating Voltage	$100\ \mu A\ I_{SET} \leq 1$ mA		0.9			0.9		V
Average Change in Set Current with Input Voltage	$100\ \mu A \leq I_{SET} \leq 1$ mA $1.5 \leq V^+ \leq 5V$ $5V \leq V^+ \leq 30V$		0.02 0.01	0.05 0.03		0.02 0.01	0.01 0.05	%/V %/V
Temperature Dependence of Set Current (Note 3) and	$100\ \mu A \leq I_{SET} \leq 1$ mA	0.98T	T	1.02T	0.97T	T	1.03T	
Equivalent Slope Error				±2			±3	%
Effective Shunt Capacitance			15			15		pF

Typical Performance Characteristics

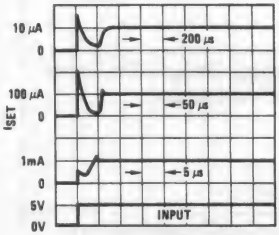
Output Impedance



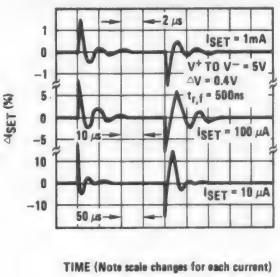
Maximum Slew Rate for Linear Operation



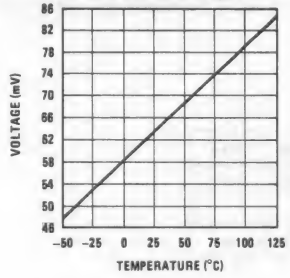
Start-Up



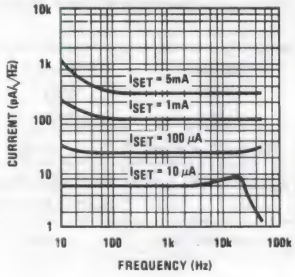
Transient Response



Voltage Across R_{SET}



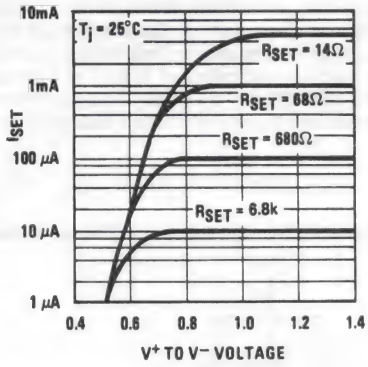
Current Noise



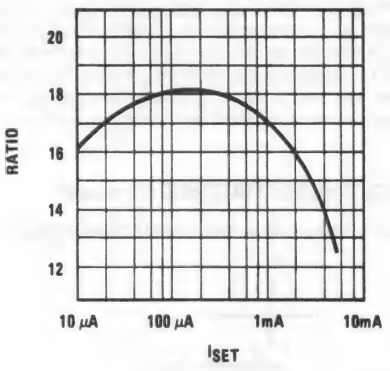
TL/H/5697–2

Typical Performance Characteristics (Continued)

Turn-On Voltage



Ratio of I_{SET} to V⁻ Current



TL/H/5697–3

Application Hints

The LM134 has been designed for ease of application, but a general discussion of design features is presented here to familiarize the designer with device characteristics which may not be immediately obvious. These include the effects of slewing, power dissipation, capacitance, noise, and contact resistance.

SLEW RATE

At slew rates above a given threshold (see curve), the LM134 may exhibit non-linear current shifts. The slewing rate at which this occurs is directly proportional to I_{SET}. At $I_{SET} = 10\ \mu A$, maximum dV/dt is $0.01V/\mu s$; at $I_{SET} = 1$ mA, the limit is $1V/\mu s$. Slew rates above the limit do not harm the LM134, or cause large currents to flow.

THERMAL EFFECTS

Internal heating can have a significant effect on current regulation for I_{SET} greater than 100 μA. For example, each 1V increase across the LM134 at $I_{SET} = 1$ mA will increase junction temperature by ≈ 0.4°C in still air. Output current (I_{SET}) has a temperature coefficient of ≈0.33%/°C, so the change in current due to temperature rise will be (0.4)(0.33) = 0.132%. This is a 10:1 degradation in regulation compared to true electrical effects. Thermal effects, therefore, must be taken into account when DC regulation is critical and I_{SET} exceeds 100 μA. Heat sinking of the TO-46 package or the TO-92 leads can reduce this effect by more than 3:1.

SHUNT CAPACITANCE

In certain applications, the 15 pF shunt capacitance of the LM134 may have to be reduced, either because of loading problems or because it limits the AC output impedance of the current source. This can be easily accomplished by buffering the LM134 with an FET as shown in the applications. This can reduce capacitance to less than 3 pF and improve regulation by at least an order of magnitude. DC characteristics (with the exception of minimum input voltage), are not affected.

NOISE

Current noise generated by the LM134 is approximately 4 times the shot noise of a transistor. If the LM134 is used as an active load for a transistor amplifier, input referred noise will be increased by about 12 dB. In many cases, this is acceptable and a single stage amplifier can be built with a voltage gain exceeding 2000.

LEAD RESISTANCE

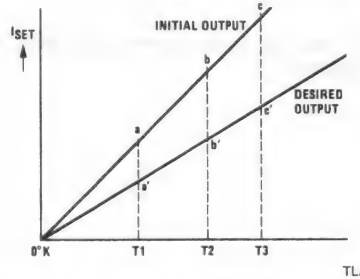
The sense voltage which determines operating current of the LM134 is less than 100 mV. At this level, thermocouple or lead resistance effects should be minimized by locating the current setting resistor physically close to the device. Sockets should be avoided if possible. It takes only 0.7Ω contact resistance to reduce output current by 1% at the 1 mA level.

SENSING TEMPERATURE

The LM134 makes an ideal remote temperature sensor because its current mode operation does not lose accuracy over long wire runs. Output current is directly proportional to absolute temperature in degrees Kelvin, according to the following formula:

$$I_{SET} = \frac{(227\mu V/°K)(T)}{R_{SET}}$$

Calibration of the LM134 is greatly simplified because of the fact that most of the initial inaccuracy is due to a gain term (slope error) and not an offset. This means that a calibration consisting of a gain adjustment only will trim both slope and zero at the same time. In addition, gain adjustment is a one point trim because the output of the LM134 extrapolates to zero at 0°K, independent of R_{SET} or any initial inaccuracy.



TL/H/5697–4

This property of the LM134 is illustrated in the accompanying graph. Line abc is the sensor current before trimming.

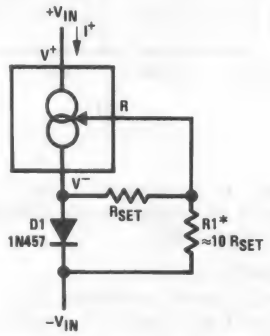
Application Hints (Continued)

Line a′b′c′ is the desired output. A gain trim done at T2 will move the output from b to b′ and will simultaneously correct the slope so that the output at T1 and T3 will be correct. This gain trim can be done on R_{SET} or on the load resistor used to terminate the LM134. Slope error after trim will normally be less than ±1%. To maintain this accuracy, however, a low temperature coefficient resistor must be used for R_{SET}.

A 33 ppm/°C drift of R_{SET} will give a 1% slope error because the resistor will normally see about the same temperature variations as the LM134. Separating R_{SET} from the LM134 requires 3 wires and has lead resistance problems, so is not normally recommended. Metal film resistors with less than 20 ppm/°C drift are readily available. Wire wound resistors may also be used where best stability is required.

Typical Applications (Continued)

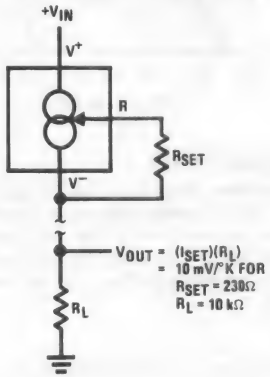
Zero Temperature Coefficient Current Source



TL/H/5697–13

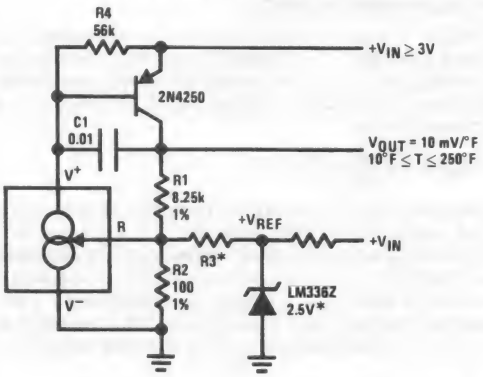
*Select ratio of R1 to R_{SET} to obtain zero drift. $I^+ \approx 2\ I_{SET}$

Terminating Remote Sensor for Voltage Output



TL/H/5697–14

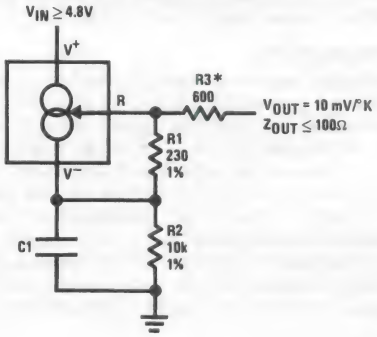
Ground Referred Fahrenheit Thermometer



TL/H/5697–15

*Select R3 = V_{REF}/583 μA. V_{REF} may be any stable positive voltage ≥2V
Trim R3 to calibrate

Low Output Impedance Thermometer

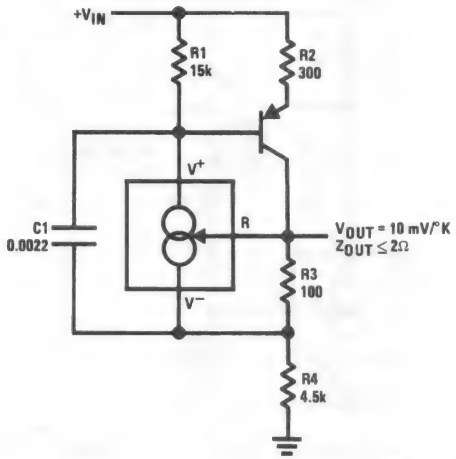


TL/H/5697–6

*Output impedance of the LM134 at the "R" pin is approximately $\frac{-R_o\Omega}{16}$ where R_o is the equivalent external resistance connected to the V⁻ pin. This negative resistance can be reduced by a factor of 5 or more by inserting an equivalent resistor in series with the output.

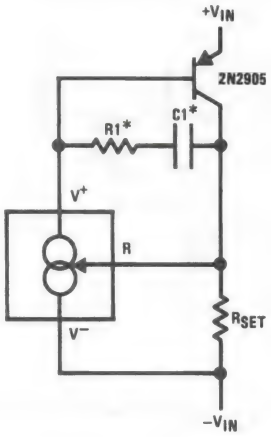
Typical Applications (Continued)

Low Output Impedance Thermometer



TL/H/5697–16

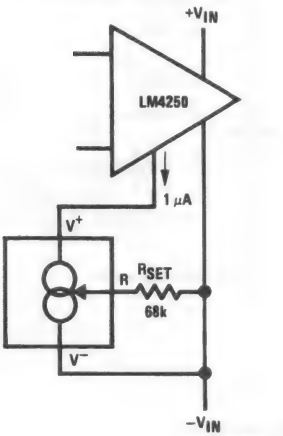
Higher Output Current



TL/H/5697–5

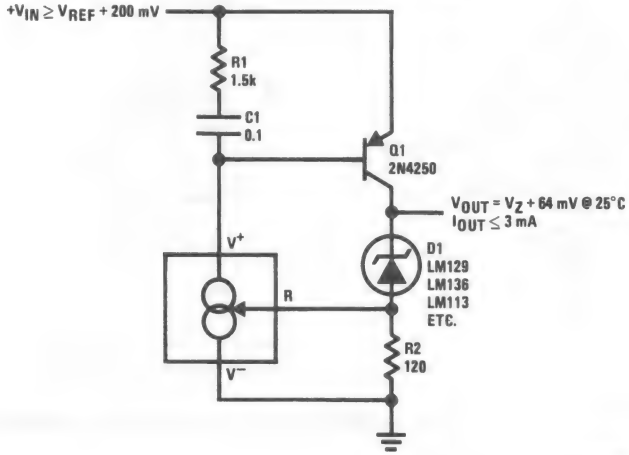
*Select R1 and C1 for optimum stability

Micropower Bias



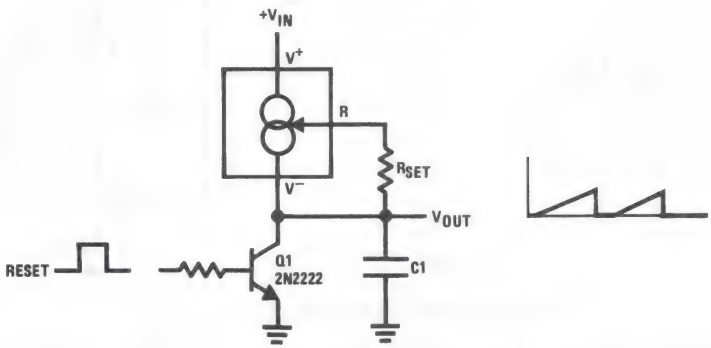
TL/H/5697–17

Low Input Voltage Reference Driver



TL/H/5697–18

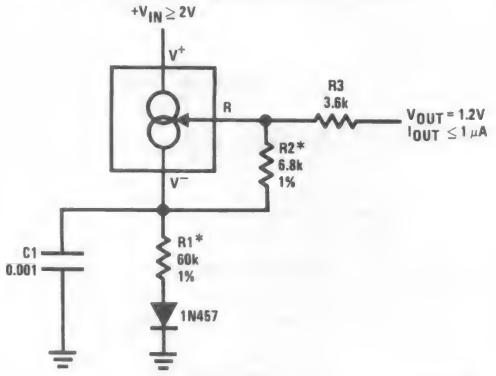
Ramp Generator



TL/H/5697–19

Typical Applications (Continued)

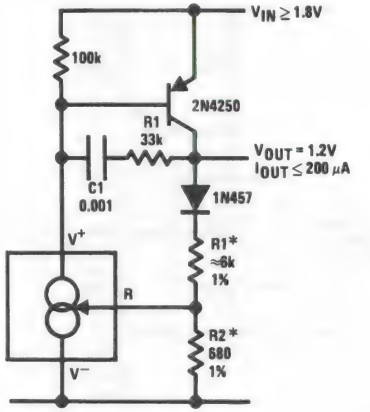
1.2V Reference Operates on 10 μA and 2V



TL/H/5697–20

*Select ratio of R1 to R2 to obtain zero temperature drift

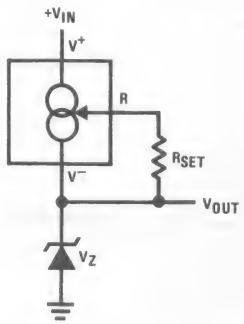
1.2V Regulator with 1.8V Minimum Input



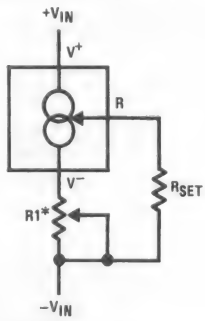
TL/H/5697–7

*Select ratio of R1 to R2 for zero temperature drift

Zener Biasing

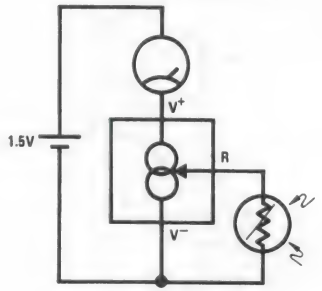


Alternate Trimming Technique



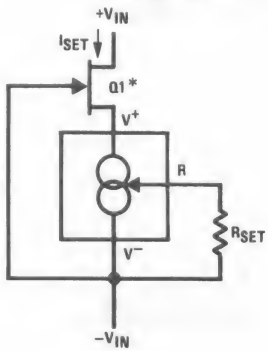
*For ±10% adjustment, select R_{SET} 10% high, and make R1 ≈ 3 R_{SET}

Buffer for Photoconductive Cell

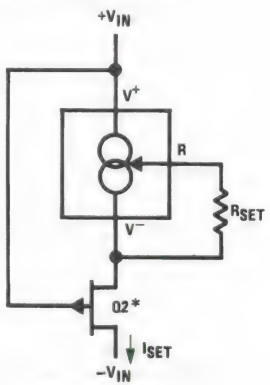


TL/H/5697–8

FET Cascoding for Low Capacitance and/or Ultra High Output Impedance



TL/H/5697–21

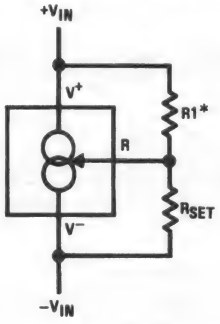


TL/H/5697–22

*Select Q1 or Q2 to ensure at least 1V across the LM134. $V_p\ (1 - I_{SET}/I_{DSS}) \geq 1.2V$.

Typical Applications (Continued)

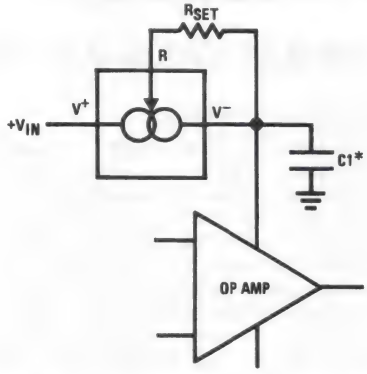
Generating Negative Output Impedance



TL/H/5697–23

*$Z_{OUT} \approx -16 \bullet R1$ ($R1/V_{IN}$ must not exceed I_{SET})

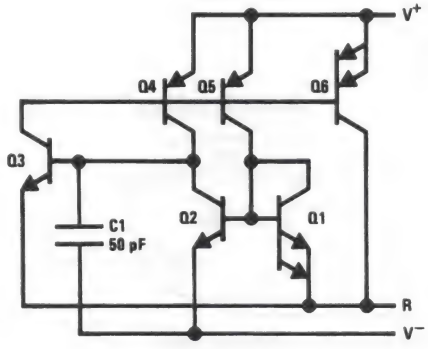
In-Line Current Limiter



TL/H/5697–9

*Use minimum value required to ensure stability of protected device. This minimizes inrush current to a direct short.

Schematic Diagram



TL/H/5697–11

National Semiconductor

LM136-2.5/LM236-2.5/LM336-2.5V Reference Diode

General Description

The LM136-2.5/LM236-2.5 and LM336-2.5 integrated circuits are precision 2.5V shunt regulator diodes. These monolithic IC voltage references operate as a low-temperature-coefficient 2.5V zener with 0.2Ω dynamic impedance. A third terminal on the LM136-2.5 allows the reference voltage and temperature coefficient to be trimmed easily.

The LM136-2.5 series is useful as a precision 2.5V low voltage reference for digital voltmeters, power supplies or op amp circuitry. The 2.5V make it convenient to obtain a stable reference from 5V logic supplies. Further, since the LM136-2.5 operates as a shunt regulator, it can be used as either a positive or negative voltage reference.

The LM136-2.5 is rated for operation over −55°C to +125°C while the LM236-2.5 is rated over a −25°C to +85°C temperature range.

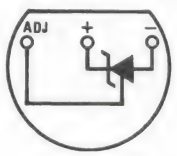
Both are packaged in a TO-46 package. The LM336-2.5 is rated for operation over a 0°C to +70°C temperature range and is available in a TO-92 plastic package.

Features

- Low temperature coefficient
- Wide operating current of 400 μA to 10 mA
- 0.2Ω dynamic impedance
- ±1% initial tolerance available
- Guaranteed temperature stability
- Easily trimmed for minimum temperature drift
- Fast turn-on
- Three lead transistor package

Connection Diagrams

TO-92
Plastic Package

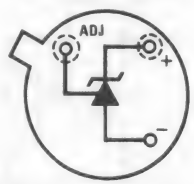


TL/H/5715–8

Bottom View

Order Number LM336Z-2.5
or LM336BZ-2.5
See NS Package Number Z03A

TO-46
Metal Can Package

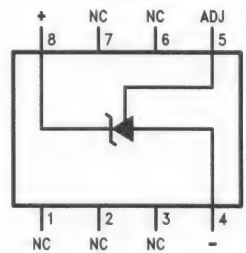


TL/H/5715–20

Bottom View

Order Number LM136H-2.5,
LM236H-2.5, LM336H-2.5,
LM136AH-2.5 or LM236AH-2.5
See NS Package Number H03H

SO Package



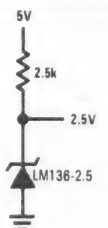
TL/H/5715–12

Top View

Order Number LM336M-2.5
or LM336BM-2.5
See NS Package Number M08A

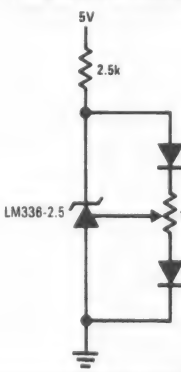
Typical Applications

2.5V Reference



TL/H/5715–9

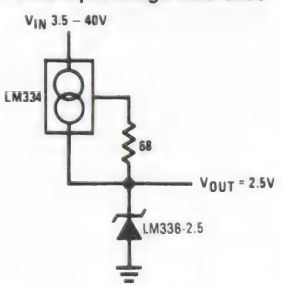
2.5V Reference with Minimum Temperature Coefficient



†Adjust to 2.490V
*Any silicon signal diode

TL/H/5715–10

Wide Input Range Reference



TL/H/5715–11

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Reverse Current	15 mA
Forward Current	10 mA
Storage Temperature	−60°C to +150°C
Operating Temperature Range	
LM136	−55°C to +150°C
LM236	−25°C to +85°C
LM336	0°C to +70°C

Soldering Information

TO-92 Package (10 sec.)	260°C
TO-46 Package (10 sec.)	300°C
SO Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (Appendix D) for other methods of soldering surface mount devices.

Electrical Characteristics (Note 1)

Parameter	Conditions	LM136A-2.5/LM236A-2.5 LM136-2.5/LM236-2.5			LM336B-2.5 LM336-2.5			Units
		Min	Typ	Max	Min	Typ	Max	
Reverse Breakdown Voltage	$T_A = 25°C$, $I_R = 1$ mA							
	LM136/LM236/LM336	2.440	2.490	2.540	2.390	2.490	2.590	V
	LM136A/LM236A, LM336B	2.465	2.490	2.515	2.440	2.490	2.540	V
Reverse Breakdown Change With Current	$T_A = 25°C$, $400\ \mu A \le I_R \le 10$ mA		2.6	6		2.6	10	mV
Reverse Dynamic Impedance	$T_A = 25°C$, $I_R = 1$ mA		0.2	0.6		0.2	1	Ω
Temperature Stability (Note 2)	V_R Adjusted to 2.490V, $I_R = 1$ mA, *(Figure 2)*							
	$0°C \le T_A \le 70°C$ (LM336)					1.8	6	mV
	$-25°C \le T_A \le +85°C$ (LM236)		3.5	9				mV
	$-55°C \le T_A \le +125°C$ (LM136)		12	18				mV
Reverse Breakdown Change With Current	$400\ \mu A \le I_R \le 10$ mA		3	10		3	12	mV
Reverse Dynamic Impedance	$I_R = 1$ mA		0.4	1		0.4	1.4	Ω
Long Term Stability	$T_A = 25°C \pm 0.1°C$, $I_R = 1$ mA		20			20		ppm

Note 1: Unless otherwise specified, the LM136-2.5 is specified from $-55°C \le T_A \le +125°C$, the LM236-2.5 from $-25°C \le T_A \le +85°C$ and the LM336-2.5 from $0°C \le T_A \le +70°C$.

Note 2: Temperature stability for the LM336 and LM236 family is guaranteed by design. Design limits are guaranteed (but not 100% production tested) over the indicated temperature and supply voltage ranges. These limits are not used to calculate outgoing quality levels. Stability is defined as the maximum change in V_{ref} from 25°C to T_A (min) or T_A (max).

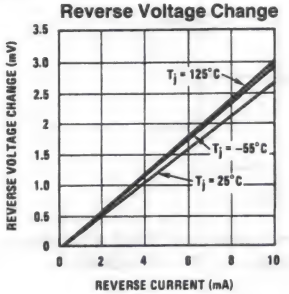
Note 3: For elevated temperature operation, T_j max is:

LM136	150°C
LM236	125°C
LM336	100°C

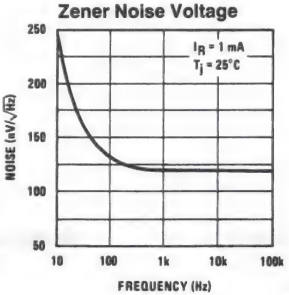
Thermal Resistance	TO-92	TO-46	SO-8
θ_{ja} (Junction to Ambient)	180°C/W (0.4" leads) 170°C/W (0.125" lead)	440°C/W	165°C/W
θ_{jc} (Junction to Case)	n/a	80°C/W	n/a

Typical Performance Characteristics

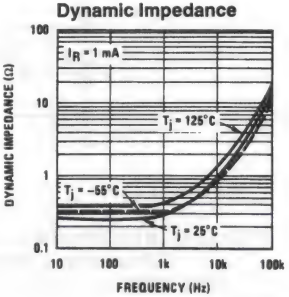
Reverse Voltage Change



Zener Noise Voltage



Dynamic Impedance



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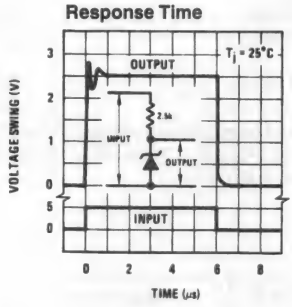
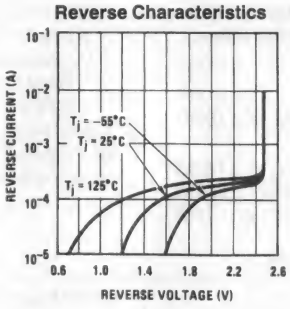
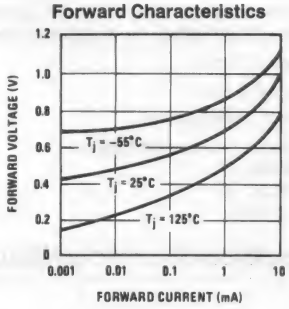
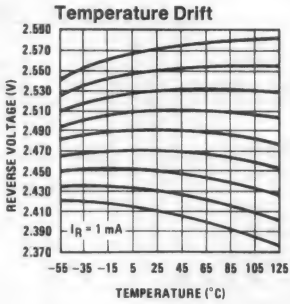
Typical Performance Characteristics (Continued)









TL/H/5715–3

Application Hints

The LM136 series voltage references are much easier to use than ordinary zener diodes. Their low impedance and wide operating current range simplify biasing in almost any circuit. Further, either the breakdown voltage or the temperature coefficient can be adjusted to optimize circuit performance.

Figure 1 shows an LM136 with a 10k potentiometer for adjusting the reverse breakdown voltage. With the addition of R1 the breakdown voltage can be adjusted without affecting the temperature coefficient of the device. The adjustment range is usually sufficient to adjust for both the initial device tolerance and inaccuracies in buffer circuitry.

If minimum temperature coefficient is desired, two diodes can be added in series with the adjustment potentiometer as shown in *Figure 2.* When the device is adjusted to 2.490V the temperature coefficient is minimized. Almost any silicon signal diode can be used for this purpose such as a 1N914, 1N4148 or a 1N457. For proper temperature compensation the diodes should be in the same thermal environment as the LM136. It is usually sufficient to mount the diodes near the LM136 on the printed circuit board. The absolute resistance of R1 is not critical and any value from 2k to 20k will work.

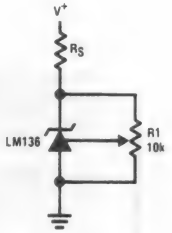
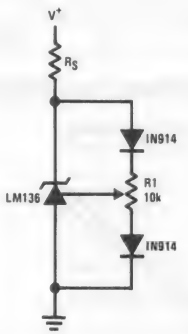


FIGURE 1. LM136 With Pot for Adjustment of Breakdown Voltage (Trim Range = ±120 mV typical)

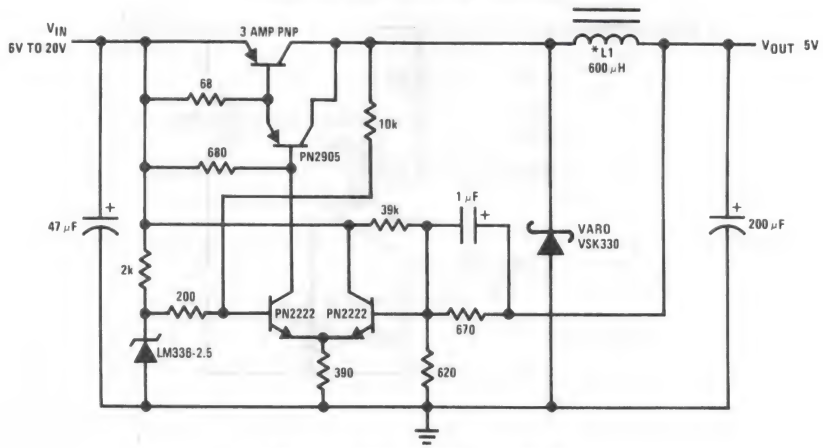


TL/H/5715–4

FIGURE 2. Temperature Coefficient Adjustment (Trim Range = ±70 mV typical)

Typical Applications (Continued)

Low Cost 2 Amp Switching Regulator†

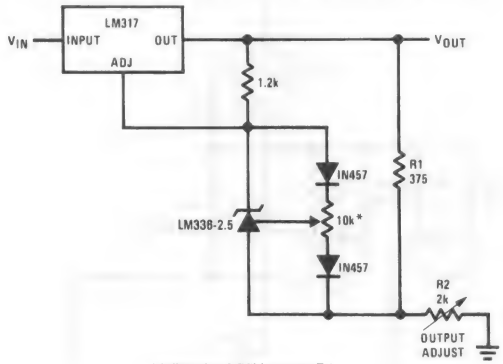


*L1 60 turns #16 wire on Arnold Core A-254168-2

†Efficiency ≈ 80%

TL/H/5715–5

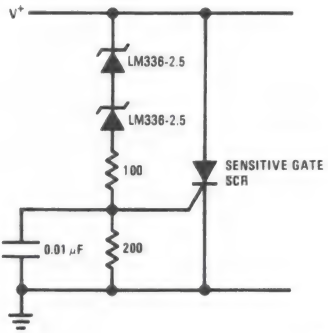
Precision Power Regulator with Low Temperature Coefficient



*Adjust for 3.75V across R1

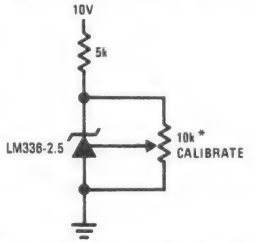
TL/H/5715–13

5V Crowbar



TL/H/5715–14

Trimmed 2.5V Reference with Temperature Coefficient Independent of Breakdown Voltage

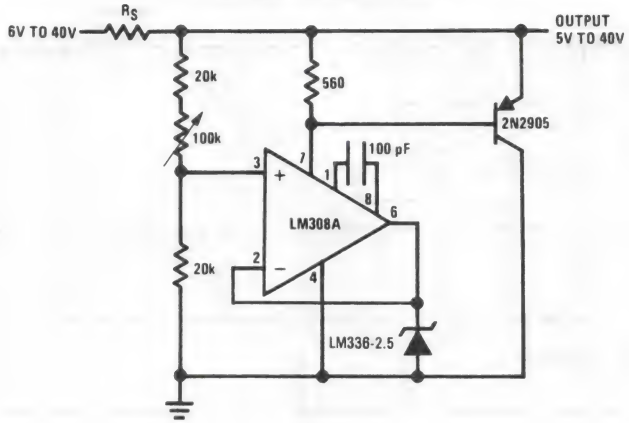


*Does not affect temperature coefficient

TL/H/5715–15

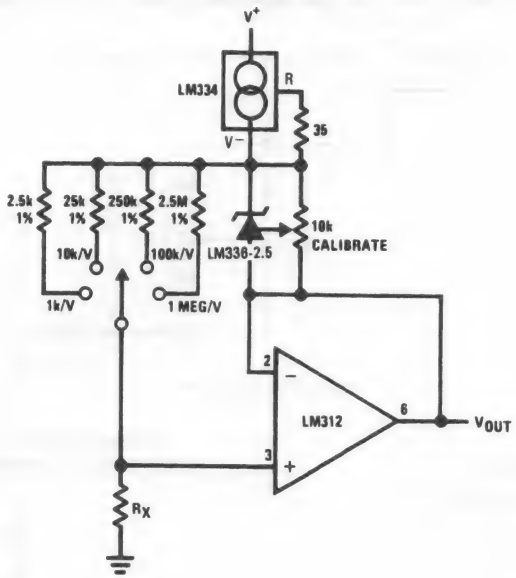
Typical Applications (Continued)

Adjustable Shunt Regulator



TL/H/5715–6

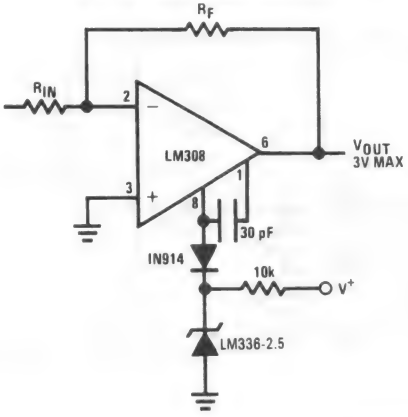
Linear Ohmmeter



TL/H/5715–16

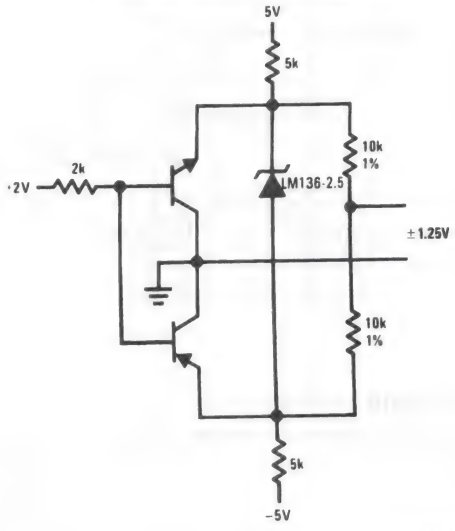
Typical Applications (Continued)

Op Amp with Output Clamped



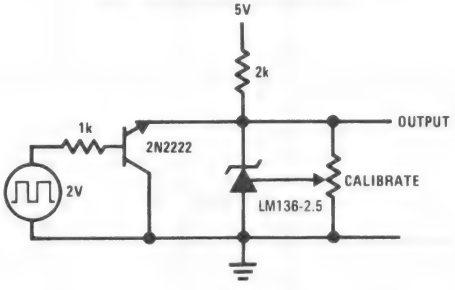
TL/H/5715–17

Bipolar Output Reference



TL/H/5715–18

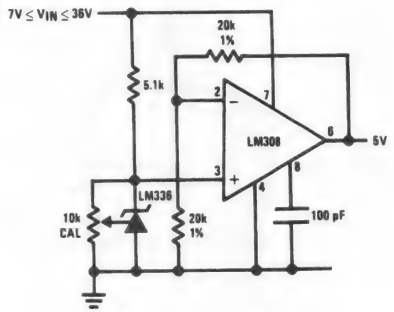
2.5V Square Wave Calibrator



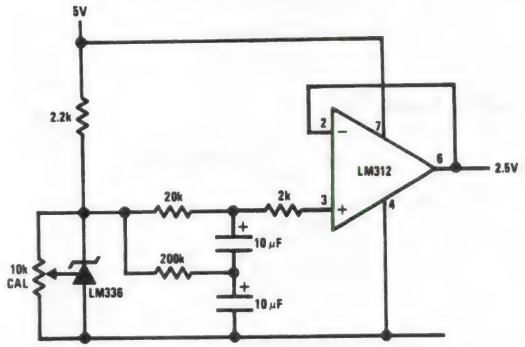
TL/H/5715–19

Typical Applications (Continued)

5V Buffered Reference

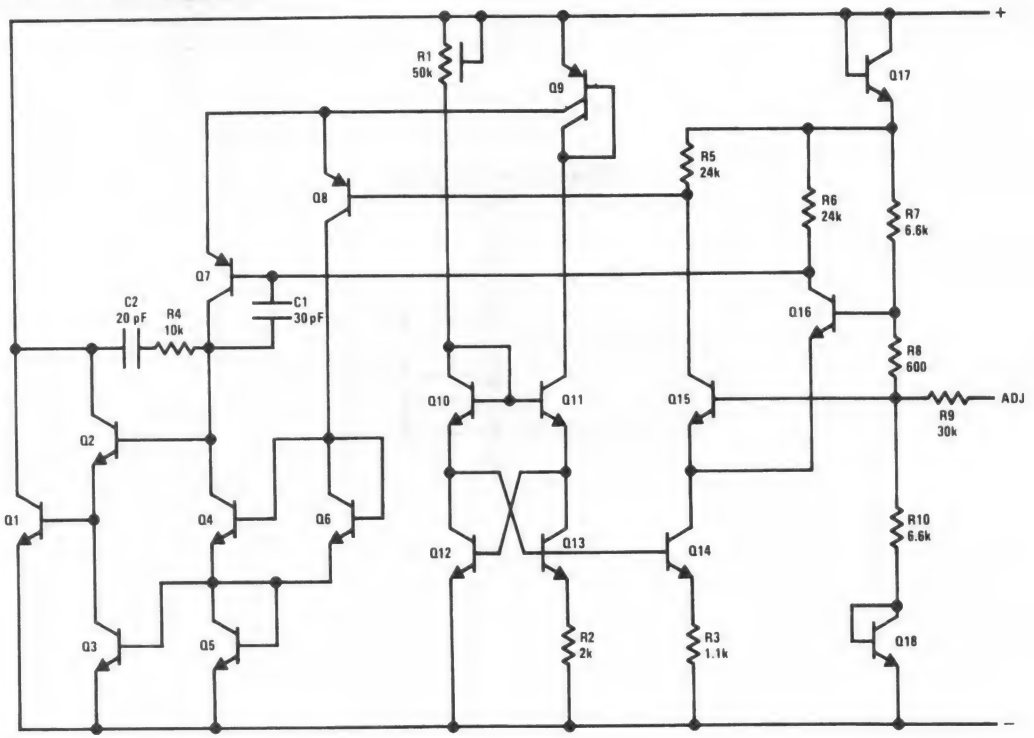


Low Noise Buffered Reference



TL/H/5715–7

Schematic Diagram



TL/H/5715–1

National Semiconductor

LM136-5.0/LM236-5.0/LM336-5.0, 5.0V Reference Diode

General Description

The LM136-5.0/LM236-5.0/LM336-5.0 integrated circuits are precision 5.0V shunt regulator diodes. These monolithic IC voltage references operate as a low temperature coefficient 5.0V zener with 0.6Ω dynamic impedance. A third terminal on the LM136-5.0 allows the reference voltage and temperature coefficient to be trimmed easily.

The LM136-5.0 series is useful as a precision 5.0V low voltage reference for digital voltmeters, power supplies or op amp circuitry. The 5.0V makes it convenient to obtain a stable reference from low voltage supplies. Further, since the LM136-5.0 operates as a shunt regulator, it can be used as either a positive or negative voltage reference.

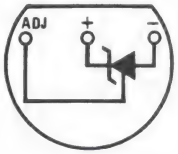
The LM136-5.0 is rated for operation over −55°C to +125°C while the LM236-5.0 is rated over a −25°C to +85°C temperature range. Both are packaged in a TO-46 package. The LM336-5.0 is rated for operation over a 0°C to +70°C temperature range and is available in a TO-92 plastic package. For applications requiring 2.5V see LM136-2.5.

Features

- Adjustable 4V to 6V
- Low temperature coefficient
- Wide operating current of 600 μA to 10 mA
- 0.6Ω dynamic impedance
- ± 1% initial tolerance available
- Guaranteed temperature stability
- Easily trimmed for minimum temperature drift
- Fast turn-on
- Three lead transistor package

Connection Diagrams

TO-92
Plastic Package

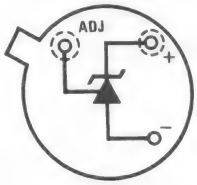


TL/H/5716–4

Bottom View

Order Number LM336Z-5.0 or LM336BZ-5.0
See NS Package Number Z03A

TO-46
Metal Can Package

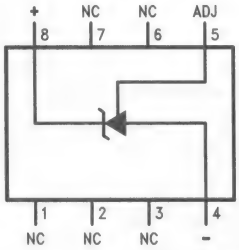


TL/H/5716–5

Bottom View

Order Number LM136H-5.0, LM236H-5.0, LM136AH-5.0 or LM236AH-5.0
See NS Package Number H03H

SO Package

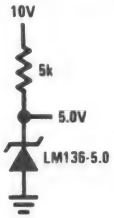


TL/H/5716–7

Order Number LM336M-5.0 or LM336BM-5.0
See NS Package Number M08A

Typical Applications

5.0V Reference



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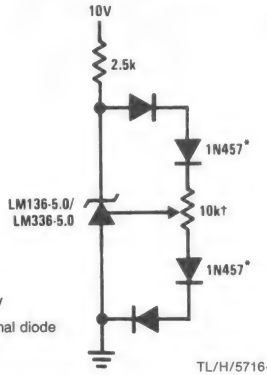
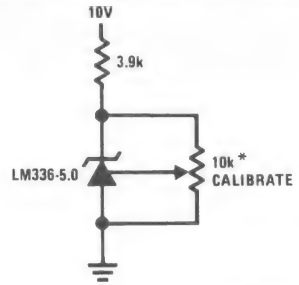
5.0V Reference with Minimum Temperature Coefficient



†Adjust to 5.00V
*Any silicon signal diode

TL/H/5716–15

Trimmed 4V to 6V Reference with Temperature Coefficient Independent of Breakdown Voltage



TL/H/5716–3

*Does not affect temperature coefficient

Absolute Maximum Ratings

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Reverse Current	15mA
Forward Current	10mA
Storage Temperature	−60°C to +150°C
Operating Temperature Range	
LM136-5.0	−55°C to +150°C
LM236-5.0	−25°C to +85°C
LM336-5.0	0°C to +70°C

Soldering Information	
TO-92 Package (10 sec.)	260°C
TO-46 Package (10 sec.)	300°C
SO Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (appendix D) for other methods of soldering surface mount devices.

Electrical Characteristics (Note 1)

Parameter	Conditions	LM136A-5.0/LM236A-5.0 LM136-5.0/LM236-5.0			LM336B-5.0 LM336-5.0			Units
		Min	Typ	Max	Min	Typ	Max	
Reverse Breakdown Voltage	$T_A = 25°C$, $I_R = 1$ mA							
	LM136-5.0/LM236-5.0/LM336-5.0	4.9	5.00	5.1	4.8	5.00	5.2	V
	LM136A-5.0/LM236A-5.0, LM336B-5.0	4.95	5.00	5.05	4.90	5.00	5.1	V
Reverse Breakdown Change With Current	$T_A = 25°C$, 600 $\mu A \le I_R \le 10$ mA		6	12		6	20	mV
Reverse Dynamic Impedance	$T_A = 25°C$, $I_R = 1$ mA		0.6	1.2		0.6	2	Ω
Temperature Stability	V_R Adjusted 5.00V, $I_R = 1$ mA, (*Figure 2*)							
	$0°C \le T_A \le 70°C$ (LM336-5.0)					4	12	mV
	$-25°C \le T_A \le +85°C$ (LM236-5.0)		7	18				mV
	$-55°C \le T_A \le +125°C$ (LM136-5.0)		20	36				mV
Reverse Breakdown Change With Current	600 $\mu A \le I_R \le 10$ mA		6	17		6	24	mV
Adjustment Range	Circuit of *Figure 1*		±1			±1		V
Reverse Dynamic Impedance	I_R = 1 mA		0.8	1.6		0.8	2.5	Ω
Long Term Stability	$T_A = 25°C \pm 0.1°C$, $I_R = 1$ mA		20			20		ppm

Note 1: Unless otherwise specified, the LM136-5.0 is specified from $-55°C \le T_A \le +125°C$, the LM236-5.0 from $-25°C \le T_A \le +85°C$ and the LM336-5.0 from $0°C \le T_A \le +70°C$.

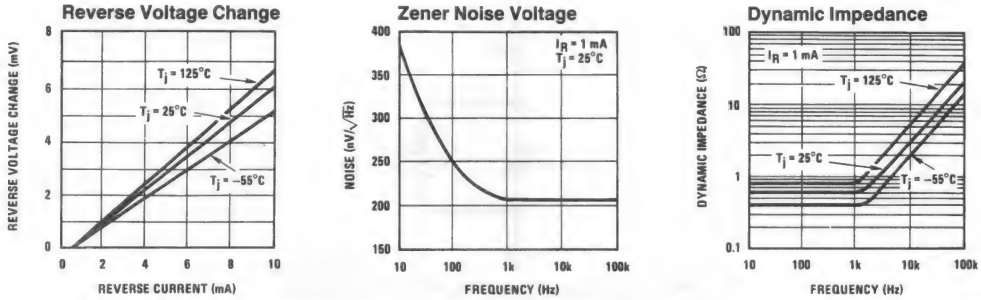
Note 2: Temperature stability for the LM336 and LM236 family is guaranteed by design. Design limits are guaranteed (but not 100% percent production tested) over the indicated temperature and supply voltage ranges. These limits are not used to calculate outgoing quality levels. Stability is defined as the maximum charge in V_{REF} from 25°C to T_A(min) or T_A(max).

Note 3: For elevated temperature operation, T_j max is:

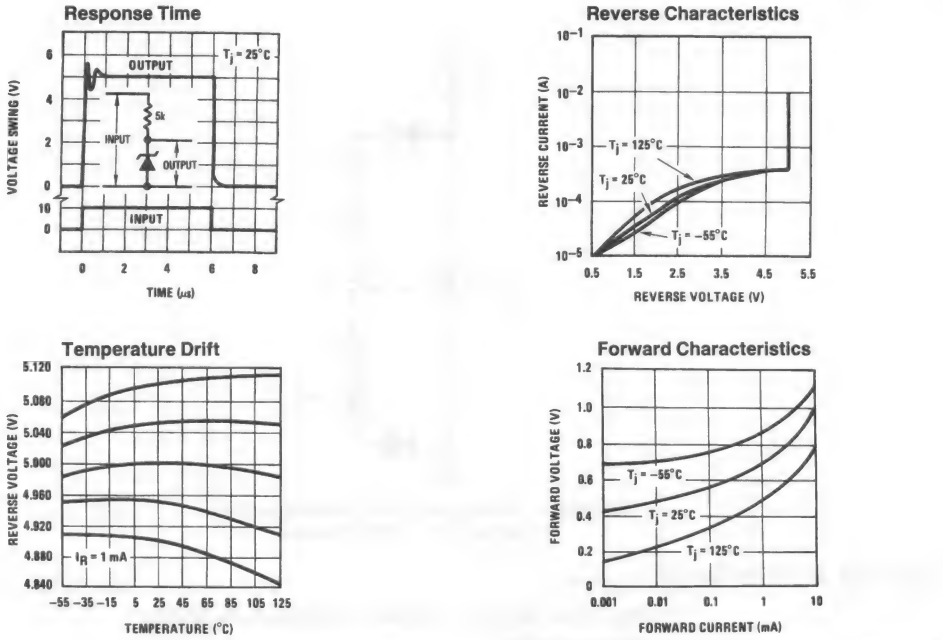
LM136 150°C
LM236 125°C
LM336 100°C

Thermal Resistance	TO-92	TO-46	SO-8
θ_{ja} (Junction to Ambient)	180°C/W (0.4" Leads) 170°C/W (0.125" Leads)	440°C/W	165°C/W
θ_{ja} (Junction to Case)	N/A	80°C/W	N/A

Typical Performance Characteristics



TL/H/5716–2



TL/H/5716–8

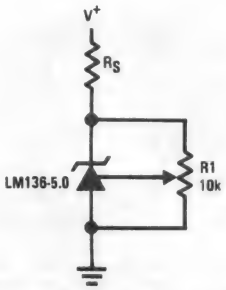
Application Hints

The LM136-5.0 series voltage references are much easier to use than ordinary zener diodes. Their low impedance and wide operating current range simplify biasing in almost any circuit. Further, either the breakdown voltage or the temperature coefficient can be adjusted to optimize circuit performance.

Figure 1 shows an LM136-5.0 with a 10k potentiometer for adjusting the reverse breakdown voltage. With the addition of R1 the breakdown voltage can be adjusted without affecting the temperature coefficient of the device. The adjustment range is usually sufficient to adjust for both the initial device tolerance and inaccuracies in buffer circuitry.

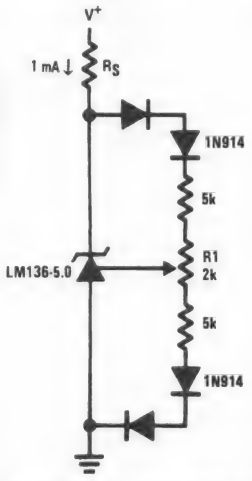
If minimum temperature coefficient is desired, four diodes can be added in series with the adjustment potentiometer as shown in *Figure 2*. When the device is adjusted to 5.00V the temperature coefficient is minimized. Almost any silicon signal diode can be used for this purpose such as a 1N914, 1N4148 or a 1N457. For proper temperature compensation the diodes should be in the same thermal environment as the LM136-5.0. It is usually sufficient to mount the diodes near the LM136-5.0 on the printed circuit board. The absolute resistance of the network is not critical and any value from 2k to 20k will work. Because of the wide adjustment range, fixed resistors should be connected in series with the pot to make pot setting less critical.

Application Hints (Continued)



TL/H/5716–9

FIGURE 1. LM136-5.0 with Pot for Adjustment of Breakdown Voltage (Trim Range = ±1.0V Typical)

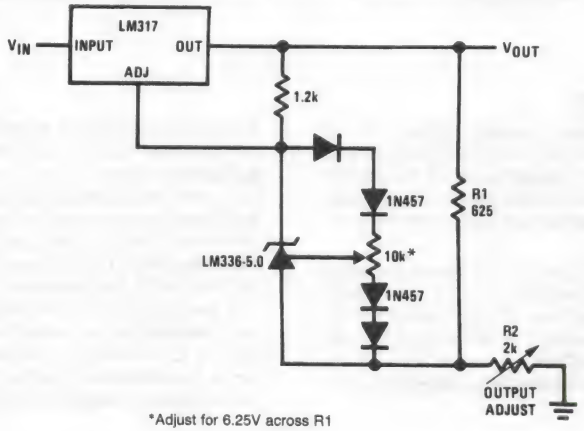


TL/H/5716–10

FIGURE 2. Temperature Coefficient Adjustment (Trim Range = ±0.5V Typical)

Typical Applications (Continued)

Precision Power Regulator with Low Temperature Coefficient

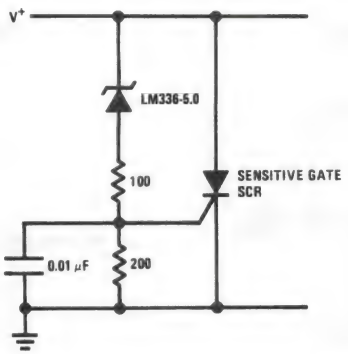


*Adjust for 6.25V across R1

TL/H/5716–11

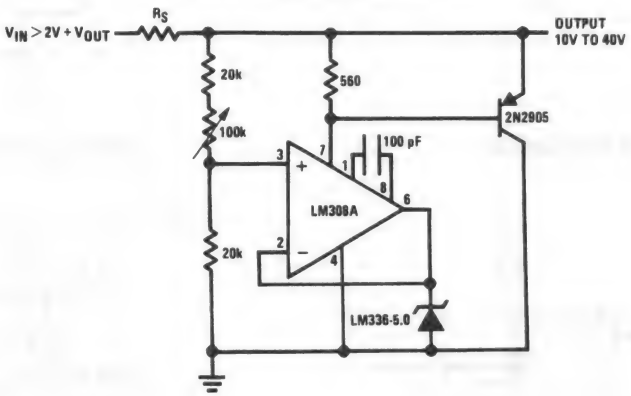
Typical Applications (Continued)

5V Crowbar



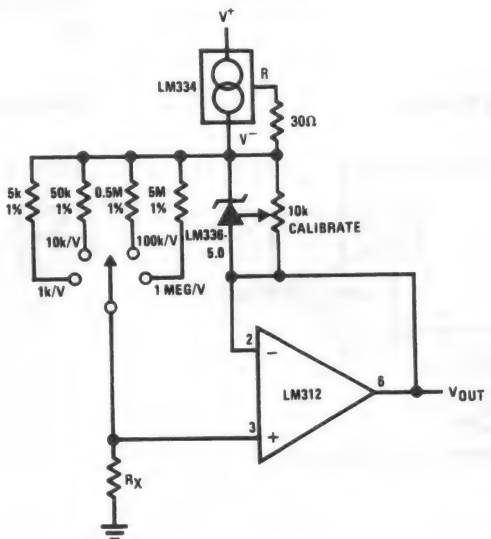
TL/H/5716–12

Adjustable Shunt Regulator



TL/H/5716–13

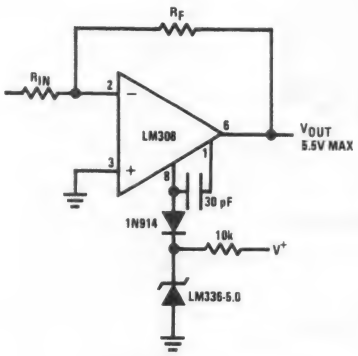
Linear Ohmmeter



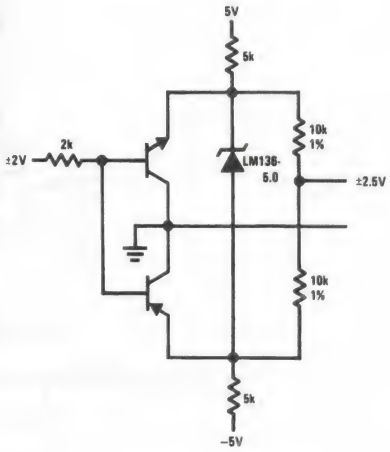
TL/H/5716–14

Typical Applications (Continued)

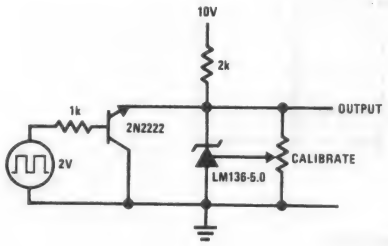
Op Amp with Output Clamped



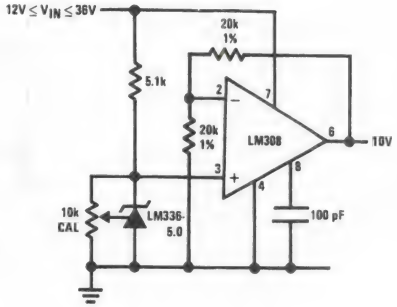
Bipolar Output Reference



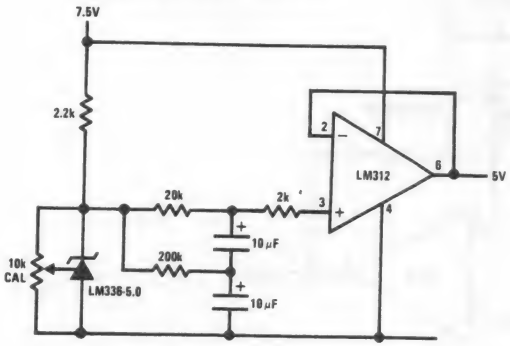
5.0V Square Wave Calibrator



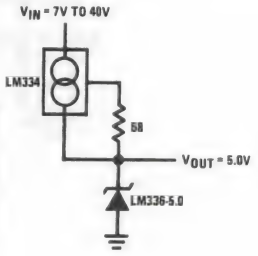
10V Buffered Reference



Low Noise Buffered Reference

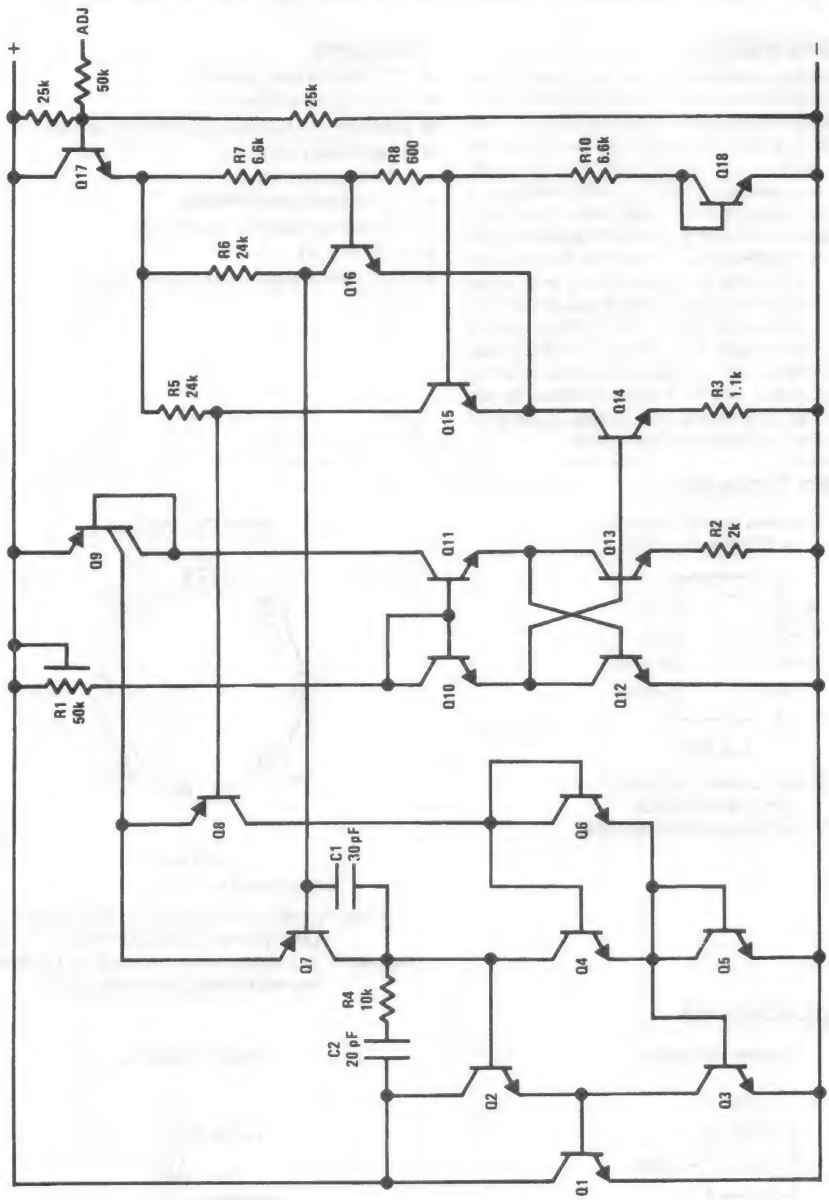


Wide Input Range Reference



TL/H/5716–6

Schematic Diagram



National Semiconductor

PRELIMINARY

LM168/LM268/LM368 Precision Voltage Reference

General Description

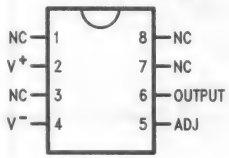
The LM168/LM368 are precision, monolithic, temperature-compensated voltage references. The LM168 makes use of thin-film technology enhanced by the discrete laser trimming of resistors to achieve excellent Temperature coefficient (Tempco) of V_{OUT} (as low as 5ppm/°C), along with tight initial tolerance, (as low as 0.02%). The trim scheme is such that individual resistors are cut open rather than being trimmed (partially cut), to avoid resistor drift caused by electromigration in the trimmed area. The LM168 also provides excellent stability vs. changes in input voltage and output current (both sourcing and sinking). This device is available in output voltage options of 5.0V and 10.0V and will operate in both series or shunt mode. Also see the LM368-2.5 data sheet for a 2.5V output. The devices are short circuit proof when sourcing current. A trim pin is made available for fine trimming of V_{OUT} or for obtaining intermediate values without greatly affecting the Tempco of the device.

Features

- 300 μA operating current
- Low output impedance
- Excellent line regulation (.0001%/V typical)
- Single-supply operation
- Externally trimmable
- Low temperature coefficient
- Operates in series or shunt mode
- 10.0V or 5.0V
- Excellent initial accuracy (0.02% typical)

Connection Diagram

Dual-In-Line Package (N)
or S.O. Package (M)

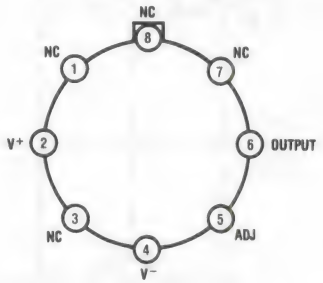


TL/H/5522–19

Top View

Order Number LM368N-5.0
or LM268BYN-5.0
See NS Package Number N08E

Metal Can Package



TL/H/5522–1

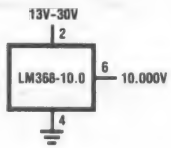
Top View

*case connected to V−

Order Number LM168BYH-10, LM168BYH-5.0,
LM268BYH-10, LM268BYH-5.0,
LM368YH-10, LM368YH-5.0, LM368H-10, LM368H-5.0
See NS Package Number H08C

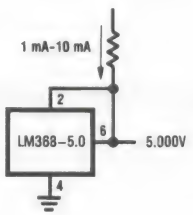
Typical Applications

Series Regulator



TL/H/5522–2

Shunt Regulator



TL/H/5522–3

Absolute Maximum Ratings (Note 8)

Input Voltage (Series Mode)	35V
Reverse Current (Shunt Mode)	50 mA
Power Dissipation	600 mW
Storage Temperature Range	−60°C to +150°C
Operating Temperature Range	
LM168	−55°C to +125°C
LM268	−40°C to +85°C
LM368	0°C to +70°C

Soldering Information

DIP (N) Package, 10 sec.	+260°C
TO-5 (H) Package, 10 sec.	+300°C
SO (M) Package, Vapor Phase (60 sec.)	+215°C
Infrared (15 sec.)	+220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (Appendix D) for other methods of soldering surface mount devices.

Electrical Characteristics (Note 1)

Parameter	Conditions	LM168/LM268/LM368			
		Typical	Tested Limit (Note 2)	Design Limit (Note 3)	Units (Max. unless noted)
V_{OUT} Error: LM168B, LM268B		±0.02	±0.05		%
LM368		±0.02	±0.1		%
Line Regulation	$(V_{OUT} + 3V) \le V_{IN} \le 30V$	±0.0001	±0.0005		%/V
Load Regulation (Note 4)	$0\ mA \le I_{SOURCE} \le 10\ mA$	±0.0003	±0.001		%/mA
	$-10\ mA \le I_{SINK} \le 0\ mA$	±0.003	±0.008		%/mA
Thermal Regulation	T = 20 mS (Note 5)	±0.005	±0.01		%/100 mW
Quiescent Current		250	350		μA
Change of Quiescent Current vs. V_{IN}	$(V_{OUT} + 3V) \le V_{IN} \le 30V$	3	5		μA/V
Temperature Coefficient of V_{OUT} (see graph): LM168BY	$-55°C \le T_A \le 125°C$	±5	±10		ppm/°C
(Note 6) LM268BY	$-40°C \le T_A \le 85°C$	±7.5	±15		ppm/°C
LM368Y	$0°C \le T_A \le 70°C$	±11	±20		ppm/°C
LM368	$0°C \le T_A \le 70°C$	±15		±30	ppm/°C
Short Circuit Current	$V_{OUT} = 0$	30	70	100	mA
Noise: 10.0V: 0.1 - 10Hz		30			uVp-p
100Hz - 10 kHz		1100			$nV/\sqrt{Hz}$
6.2V: 0.1 - 10Hz		20			uVp-p
100Hz - 10 kHz		700			$nV/\sqrt{Hz}$
5.0V: 0.1 - 10Hz		16			uVp-p
100Hz - 10 kHz		575			$nV/\sqrt{Hz}$
V_{OUT} Adjust Range: 10.000V	$0V \le V_{PIN5} \le V_{OUT}$	4.5-17.0		6.0-15.5	V min.
5.000V		4.4-7.0		4.5-6.0	V min.

Note 1: Unless otherwise noted, these specifications apply: $T_A = 25°C$, $V_{IN} = 15V$, $I_{LOAD} = 0$, $0 \le C_L \le 200$ pF, Circuit is operating in Series Mode. Or, circuit is operating in Shunt Mode, $V_{IN} = +15V$ or $V_{IN} = V_{OUT}$, $TA = +25°C$, $I_{LOAD} = -1.0$ mA, $0 \le C_L \le 200$ pF.

Note 2: Tested Limits are guaranteed and 100% tested in production.

Note 3: Design Limits are guaranteed (but not 100% production tested) over the indicated temperature and supply voltage ranges. These limits are not used to calculate outgoing quality levels.

Note 4: The LM168 has a Class B output, and will exhibit transients at the crossover point. This point occurs when the device is asked to sink approximately 120 μA. In some applications it may be advantageous to preload the output to either V_{IN} or Ground, to avoid this crossover point.

Note 5: Thermal Regulation is defined as the change in the output Voltage at a time T after a step change in power dissipation of 100 mW.

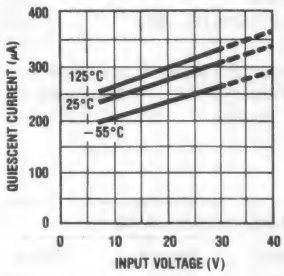
Note 6: Temperature Coefficient of V_{OUT} is defined as the worst case delta-V_{OUT} measured at Specified Temperatures divided by the total span of the Specified Temperature Range (See graphs). There is no guarantee that the Specified Temperatures are exactly at the minimum or maximum deviation.

Note 7: In metal can (H), θ_{J-C} is 75°C/W and θ_{J-A} is 150°C/W. In plastic DIP, θ_{J-A} is 160°C/W. In SO-8, θ_{J-A} is 180°C/W, in TO-92, θ_{J-A} is 160°C/W.

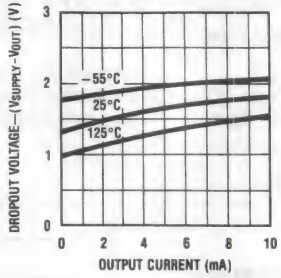
Note 8: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its Rated Operating Conditions (see Note 1 and Conditions).

Typical Performance Characteristics (Note 1)

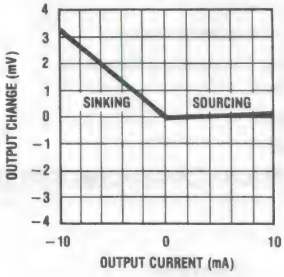
Quiescent Current vs. Input Voltage and Temperature



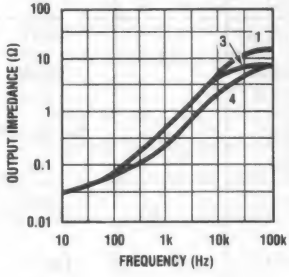
Dropout Voltage vs. Output Current (Series Mode Sourcing Current)



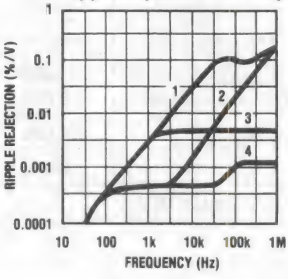
Output Change vs. Output Current



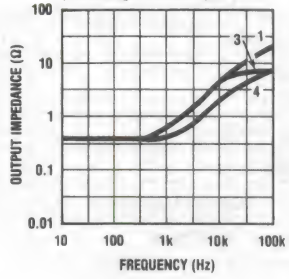
Output Impedance vs. Frequency (Sourcing Current)



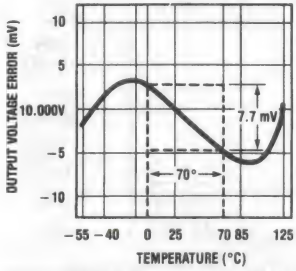
Ripple Rejection vs. Frequency



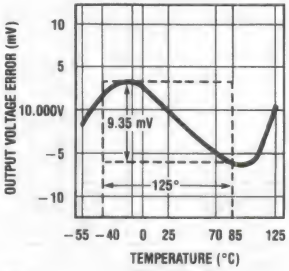
Output Impedance vs. Frequency (Sinking Current)



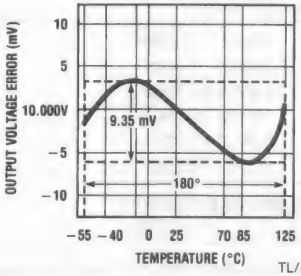
Temperature Coefficient: LM368-10 (Curve A)



Temperature Coefficient: LM268-10 (Curve B)



Temperature Coefficient: LM168-10 (Curve C)



TL/H/5522–4

Typical Temperature Coefficient Calculations:

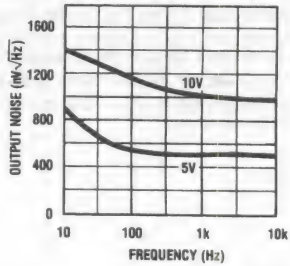
LM368-10 (see Curve A)
T.C. = 7.7 mV/(70° × 10V)
= 11 × 10E-6 = 11ppm/°C

LM268-10 (see Curve B)
T.C. = 9.35 mV/(125° × 10V)
= 7.5 × 10E-6 = 7.5ppm/°C

LM168-10 (see Curve C)
T.C. = 9.35 mV/(180° × 10V)
= 5.2 × 10E-6 = 5.2ppm/°C

(1) LM368 alone.
(2) with 0.01 μf Mylar, Trim to Gnd.
(3) with 10Ω in series with 10 μf, V_{OUT} to Gnd.
(4) with Both.

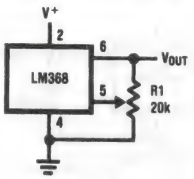
Output Noise vs. Frequency



TL/H/5522–5

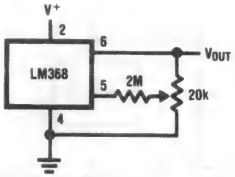
Typical Applications

Wide Range Trimmable Regulator



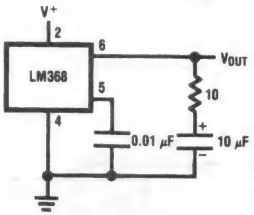
TL/H/5522–7

Narrow Range Trimmable Regulator (±1% min.)



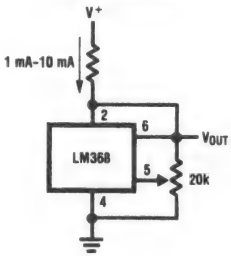
TL/H/5522–8

Improved Noise Performance



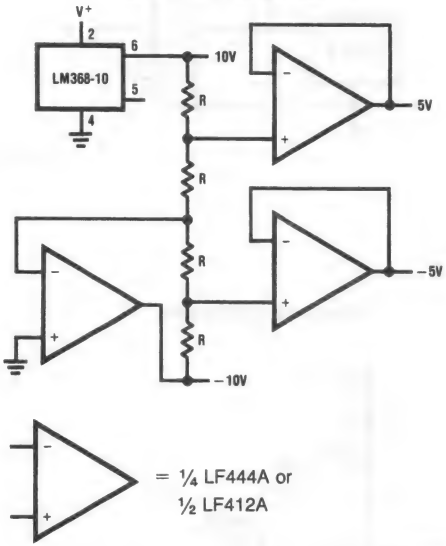
TL/H/5522–10

Adjustable Zener



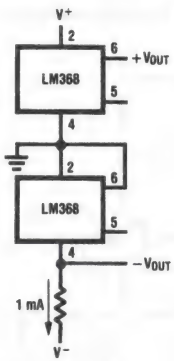
TL/H/5522–9

± 10V, ±5V References



TL/H/5522–12

± Reference

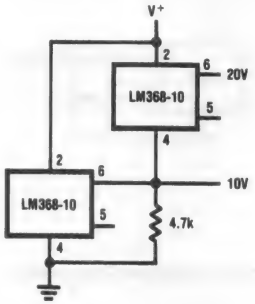


TL/H/5522–11

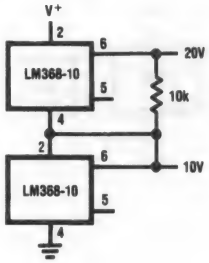
R= Thin Film Resistor Network,
±0.05% Matching and 5ppm Tracking
(Beckman 694-3-R-10K-A),
(Caddock T-914-10K-100-05)
or similar.

Typical Applications (Continued)

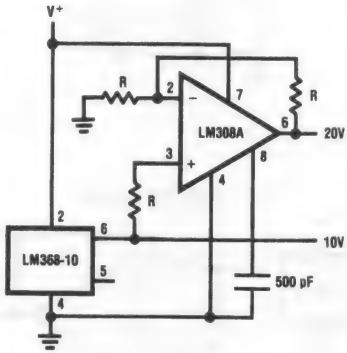
Multiple Output Voltages



TL/H/5522–13



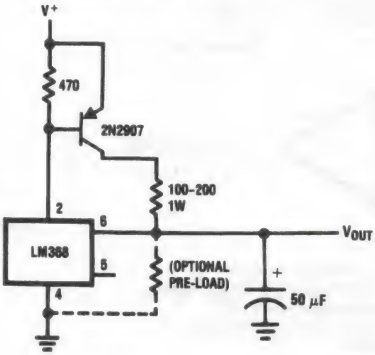
TL/H/5522–14



TL/H/5522–15

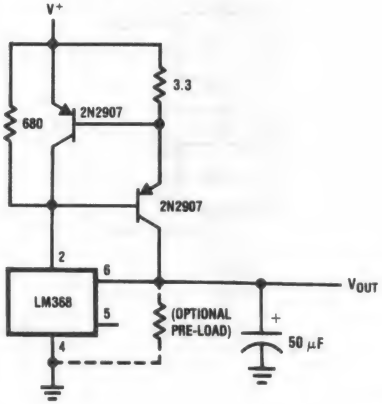
R= Thin Film Resistor Network
0.05% Matching and 5ppm Tracking
(Beckman 694-3-R-10K-A),
(Caddock T-914-10K-100-05)
or similar.

Reference with Booster



TL/H/5522–16

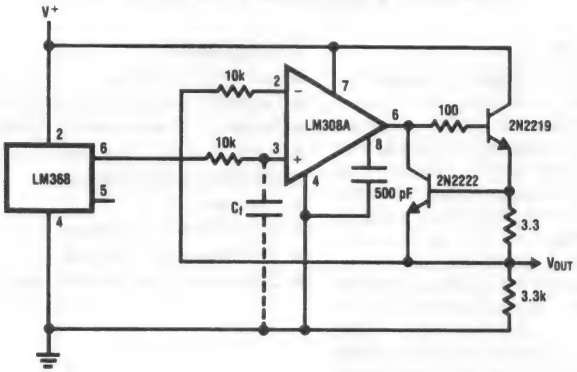
100 mA Boosted Reference



TL/H/5522–17

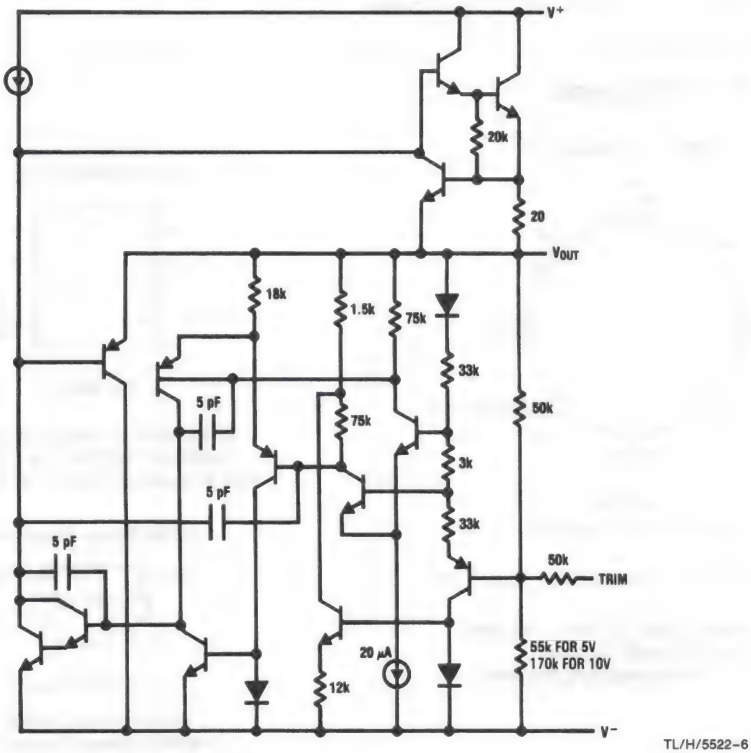
Typical Applications (Continued)

Buffered High-Current Reference with Filter



TL/H/5522–18

Simplified Schematic Diagram



TL/H/5522–6

*Reg. U.S. Pat. Off.

National Semiconductor

PRELIMINARY

LM169/LM369 Precision Voltage Reference

General Description

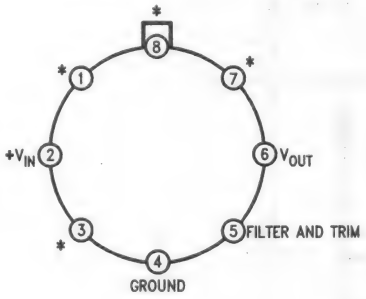
The LM169/LM369 are precision monolithic temperature-compensated voltage references. They are based on a buried zener reference as pioneered in the LM199 references, but do not require any heater, as they rely on special temperature-compensation techniques (Patent Pending). The LM169 makes use of thin-film technology enhanced by the discrete laser trimming of resistors to achieve excellent Temperature coefficient (Tempco) of V_{out} (as low as 1 ppm/°C), along with tight initial tolerances (as low as 0.01%). The trim scheme is such that individual resistors are cut open rather than being trimmed (partially cut), to avoid resistor drift caused by electromigration in the trimmed area. The LM169 also provides excellent stability vs. changes in input voltage and output current (both sourcing and sinking). The devices have a 10.000V output and will operate in either series or shunt mode; the output is short-circuit-proof to ground. A trim pin is available which permits fine-trimming of V_{out}, and also permits filtering to greatly decrease the output noise by adding a small capacitor (0.05 to 0.5 μF).

Features

- Low Tempco of V_{out}
- Excellent initial accuracy (0.003%)
- Excellent line regulation (2 ppm/V)
- Excellent output impedance
- Excellent thermal regulation
- Low noise
- Easy to filter output noise
- Low dissipation — 20 mW
- Operates in series or shunt mode

Connection Diagrams

Metal Can Package (H)



TL/H/9110–1

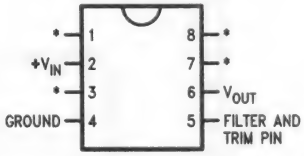
Top View

(Case is connected to ground.)

*Do not connect; internal connection for factory trims.

Order Number LM169H, LM169BH, LM369H, LM369BH,
See NS Package Number H08C

Dual-In-Line Package (N) or S.O. Package (M)



TL/H/9110–5

Top View

Order Number LM369DM, LM369N, LM369BN, LM369CN or LM369DN
See NS Package Number M08A or N08E

TO-92 Plastic Package (Z)

TL/H/9110–28

Bottom View

Order Number LM369DZ
See NS Package Number Z03A

Absolute Maximum Ratings (Note 8)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Input Voltage (Series Mode)	35V
Reverse Current (Shunt Mode)	50 mA
Power Dissipation (Note 7)	600 mW
Storage Temperature Range	−60°C to +150°C
Operating Temperature Range	(T_j min to T_j max)
LM169	−55°C to +125°C
LM369	0°C to +70°C

Soldering Information

DIP (N) Package, 10 sec.	+260°C
TO-5 (H) Package, 10 sec.	+300°C
SO (M) Package, Vapor Phase (60 sec.)	+215°C
Infrared (15 sec.)	+220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (Appendix D) for other methods of soldering surface mount devices.

ESD Tolerance

C_{zap} = 100 pF, R_{zap} = 1.5k 800V

Electrical Characteristics, LM169, LM369 (Note 1)

Parameter	Conditions	Typical	Tested Limits (Note 2)	Design Limit (Note 3)	Units (Max Unless Noted)
V_{out} Nominal		+10.000			V
V_{out} Error	(Note 11)	50 0.50	±500 ±5		ppm mV
V_{out} Tempco LM169B, LM369B LM169, LM369 LM369C (Note 6) (Note 11)	 $T_{min} < T_j < T_{max}$ $T_{min} < T_j < T_{max}$ $T_{min} < T_j < T_{max}$	 1.5 2.7 6	 **3.0** **5.0** **10**	 — — —	 ppm/°C ppm/°C ppm/°C
Line Regulation	13V ≤ V_{IN} ≤ 30V	2.0	4.0	**8.0**	ppm/V
Load Regulation Sourcing Sinking (Note 12) (Note 4, Note 9)	 0 to 10 mA 0 to −10 mA	 +3 +80	 ±8.0 +150	 **20.0**	 ppm/mA ppm/mA
Thermal Regulation Sourcing Sinking (Note 12) (Note 5)	(t = 10 msec After Load is Applied)	 3.0 3.0	 ±20 —	 — —	 ppm/100 mW ppm/100 mW
Supply Current		1.4	1.8	**2.0**	mA
ΔSupply Current	13V ≤ V_{IN} ≤ 30V	0.06	0.12	**0.2**	mA
Short Circuit Current		27	15 50	**11** **65**	mA min mA max
Noise Voltage	10 Hz to 1 kHz 0.1 Hz to 10 Hz (10 Hz to 10 kHz, C_{filter} = 0.1 μF)	10 4 4	30 — —	— — —	μV rms μV p-p μV rms
Long-term Stability (Non-Cumulative) (Note 10)	1000 hours, $T_j < T_{max}$ (Measured at +25°C)	6	—	—	ppm
Temperature Hysteresis of V_{out}	ΔT = 25°C	3	—	—	ppm
Output Shift per 1 μA at Pin 5		1500	2600	—	ppm

Electrical Characteristics LM369D (Note 1)

Parameter	Conditions	Typical	Tested Limits (Note 2)	Design Limit (Note 3)	Units (Max Unless Noted)
V_{out} Nominal		+10.000			V
V_{out} Error, LM369D		70 0.7	±1000 ±10.0	— —	ppm mV
V_{out} Tempco (Note 6)	$T_{min} \le T_j \le T_{max}$	5		**30**	ppm/°C
Line Regulation	$13V \le V_{IN} \le 30V$	2.4	±6.0	**12**	ppm/V
Load Regulation Sourcing Sinking (Note 12) (Note 4, Note 9)	 0 to 10 mA 0 to −10 mA	 +3 +80	 ±12 +160	 **±25**	 ppm/mA ppm/mA
Thermal Regulation Sourcing Sinking (Note 12) (Note 5)	(t = 10 msec After Load is Applied)	 4.0 4.0	 ±25 —	 — —	 ppm/100 mW ppm/100 mW
Supply Current		1.5	2.0	**2.4**	mA
ΔSupply Current	$13V \le V_{IN} \le 30V$	0.06	0.16	**0.3**	mA
Short Circuit Current		27	14 50	**10** **65**	mA min mA max
Noise Voltage	10 Hz to 1 kHz 0.1 Hz to 10 Hz (10 Hz to 10 kHz, C_{filter} = 0.1 μF)	10 4 4	30 — —	— — —	μV rms μV p-p μV rms
Long-Term Stability (Non-Cumulative)	1000 Hours, $T_j < T_{max}$ (Measured at +25°C)	8	—	—	ppm
Temperature Hysteresis of V_{out}	ΔT = 25°C	5	—	—	ppm
Output Shift Per 1 μA at Pin 5		1500	2800	—	ppm

Note 1: Unless otherwise noted, these conditions apply: T_j = +25°C, $13V \le V_{in} \le 17V$, $0 \le I_{load} \le 1.0$ mA, $C_L = \le 200$ pF. Specifications in **BOLDFACED TYPE** apply over the rated operating temperature range.

Note 2: Tested limits are guaranteed and 100% tested in production.

Note 3: Design Limits are guaranteed (but not 100% production tested) over the indicated temperature and supply voltage ranges. These limits are not to be used to calculate outgoing quality levels.

Note 4: The LM169 has a Class B output, and will exhibit transients at the crossover point. This point occurs when the device is required to sink approximately 1.0 mA. In some applications it may be advantageous to pre-load the output to either V_{in} or to ground, to avoid this crossover point.

Note 5: Thermal regulation is defined as the change in the output voltage at a time T after a step change of power dissipation of 100 mW.

Note 6: Temperature Coefficient of V_{OUT} is defined as the worst-case ΔV_{out} measured at Specified Temperatures divided by the total span of the Specified Temperature Range (see graphs). There is no guarantee that the Specified Temperatures are exactly at the minimum or maximum deviation.

Note 7: In metal can (H), $\theta_{J\text{-}C}$ is 75°C/W and $\theta_{J\text{-}A}$ is 150°C/W. In plastic DIP, $\theta_{J\text{-}A}$ is 160°C/W. In S0-8, $\theta_{J\text{-}A}$ is 180°C/W, in TO-92, $\theta_{J\text{-}A}$ is 160°C/W.

Note 8: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications are not guaranteed beyond the Rated Operating Conditions.

Note 9: Regulation is measured at constant temperature using pulse testing with a low duty cycle. Changes in output voltage due to heating effects are covered under the specifications for Thermal Regulation and Tempco. Load Regulation is measured at a point on the output pin 1/8″ below the bottom of the package.

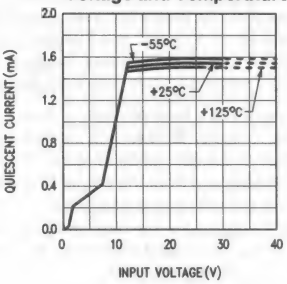
Note 10: Consult factory for availability of devices with Guaranteed Long-term Stability.

Note 11: Consult factory for availability of devices with tighter Accuracy and Tempco Specifications.

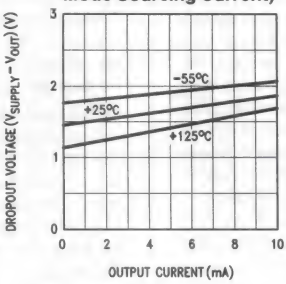
Note 12: In Sinking mode, connect 0.1 μF tantalum capacitor from output to ground.

Typical Performance Characteristics (Note 1)

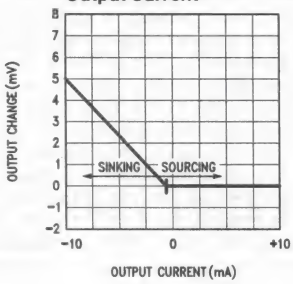
Quiescent Current vs Input Voltage and Temperature



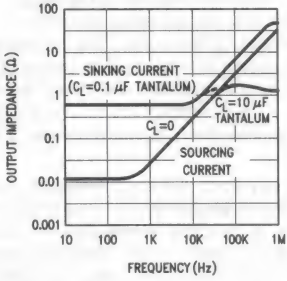
Dropout Voltage vs Output Current (Series Mode Sourcing Current)



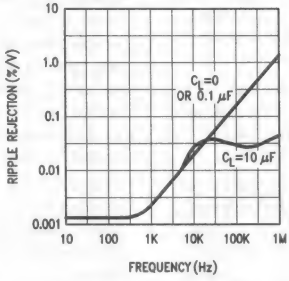
Output Change vs Output Current



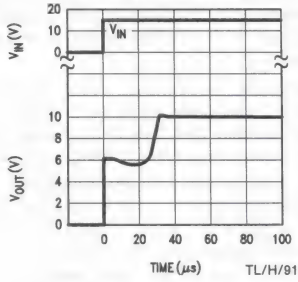
Output Impedance vs Frequency



Ripple Rejection vs Frequency

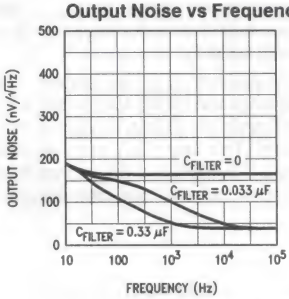


Start-up Response



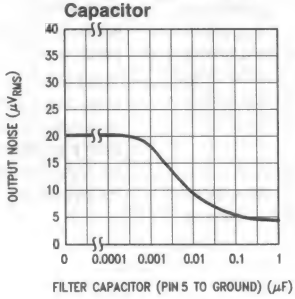
TL/H/9110–6

Output Noise vs Frequency



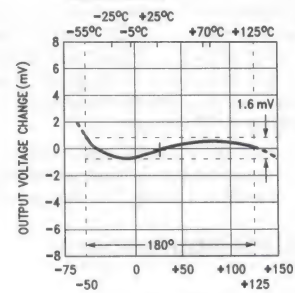
TL/H/9110–24

Output Noise vs Filter Capacitor



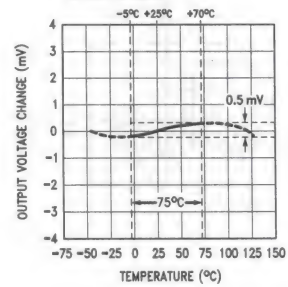
TL/H/9110–25

LM169 Temperature Coefficient



TL/H/9110–26

LM369 Temperature Coefficient



TL/H/9110–27

Typical Temperature Coefficient Calculations:

LM169 (see curve above):
T.C. = 1.6 mV/(180° × 10V)
= 8.9×10^{-7} = 0.89 ppm/°C

LM369 (see curve at left):
T.C. = 0.5 mV/(75° × 10V)
= 6.7×10^{-7} = 0.67 ppm/°C

7

Application Hints

The LM169/LM369 can be applied in the same way as any other voltage reference. The adjacent Typical Applications Circuits suggest various uses for the LM169/LM369. The LM169 is recommended for applications where the highest stability and lowest noise is required over the full military temperature range. The LM369 is suitable for limited-temperature operation. The curves showing the Noise vs. Capacitance in the Typical Performance Characteristics section show graphically that a modest capacitance of 0.1 to 0.3 microfarads can cut the broadband noise down to a level of only a few microvolts, less than 1 ppm of the output voltage. The capacitor used should be a low-leakage type. For the temperature range 0 to 50°C, polyester or Mylar® will be suitable, but at higher temperatures, a premium film capacitor such as polypropylene is recommended. For operation at +125°C, a Teflon® capacitor would be required, to ensure sufficiently low leakage. Ceramic capacitors may seem to do the job, but are not recommended for production use, as the high-K ceramics cannot be guaranteed for low leakage, and may exhibit piezo-electric effects, converting vibration or mechanical stress into excessive electrical noise.

Additionally, the inherent superiority of the LM169/369's buried Zener diode provides freedom from low-frequency noise, wobble, and jitter, in the frequency range 0.01 to 10 Hertz, where capacitive filtering is not feasible.

Pins 1, 3, 7, and 8 of the LM169/369 are connected to internal trim circuits which are used to trim the device's output voltage and Tempco during final testing at the factory. Do not connect anything to these pins, or improper operation may result. These pins would not be damaged by a short to ground, or by Electrostatic Discharges; however, keep them away from large transients or AC signals, as stray capacitance could couple noises into the output. These pins may be cut off if desired. Alternatively, a shield foil can be laid out on the printed circuit board, surrounding these pins and pin 5, and this guard foil can be connected to ground or to V_{out}, effectively acting as a guard against AC coupling and DC leakages.

The trim pin (pin 5) should also be guarded away from noise signals and leakages, as it has a sensitivity of 15 millivolts of ΔV_{out} per microampere. The trim pin can also be used in the circuits shown, to provide an output trim range of ±10 millivolts. Trimming to a wider range is possible, but is not recommended as it may degrade the Tempco and the Tempco linearity at temperature extremes. For example, if the output were trimmed up to 10.240V, the Tempco would be degraded by 8 ppm/°C. As a general rule, Tempco will be degraded by 1 ppm/°C per 30 mV of output adjustment.

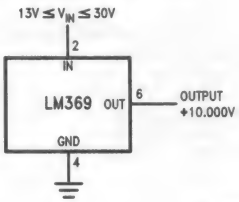
The output can sink current as well as source it, but the output impedance is much better for sourcing current. Also, the LM169/369 requires a 0.1 μF tantalum capacitor (or, 0.1 μF in series with 10Ω) bypass from the output to ground, for stable operation in shunt mode (output sinking current). The output has a class-B stage, so if the load current changes from sourcing to sinking, an output transient will occur. To avoid this transient, it may be advisable to preload the output with a few milliamperes of load to ground. The LM169/369 does have an excellent tolerance of load capacitance, and in cases of load transients, electrolytic or tantalum capacitors in the range 1 to 500 microfarads have been shown to improve the output impedance without degrading the dynamic stability of the device. The LM169/369 are rated to drive an output of ±10 mA, but for best accuracy, any load current larger than 1 mA can cause thermal errors (such as, 1 mA × 5V × 4 ppm/100 mW = 0.2 ppm or 2 microvolts) and degrade the ultimate precision of the output voltage.

The output is short-circuit-proof to ground. However, avoid overloads at high ambient temperatures, as a prolonged short-circuit may cause the junction temperature to exceed the Absolute Maximum Temperature. The device does not include a thermal shut-down circuit. If the output is pulled to a positive voltage such as +15 or +20V, the output current will be limited, but overheating may occur. Avoid such overloads for voltages higher than +20 V, for more than 5 seconds, or, at high ambient temperatures.

The LM169/369 has an excellent long-term stability, and is suitable for use in high-resolution Digital Voltmeters or Data Acquisition systems. Its long-term stability is typically 3 to 10 ppm per 1000 hours when held near T_{max}, and slightly better when operated at room temperature. Contact the factory for availability of devices with proven long-term stability.

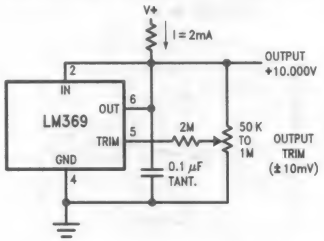
Typical Applications

Series Reference



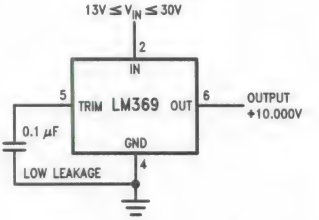
TL/H/9110–2

Shunt Reference with Optional Trim



TL/H/9110–3

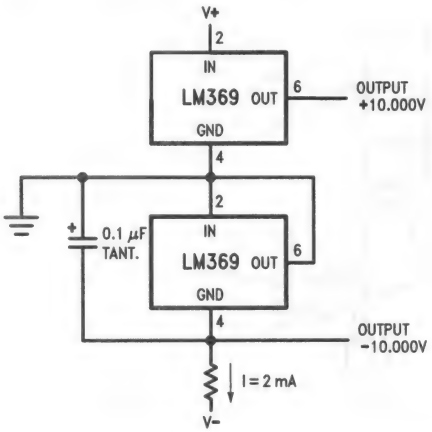
Series Reference with Optional Filter for Reduced Noise



TL/H/9110–4

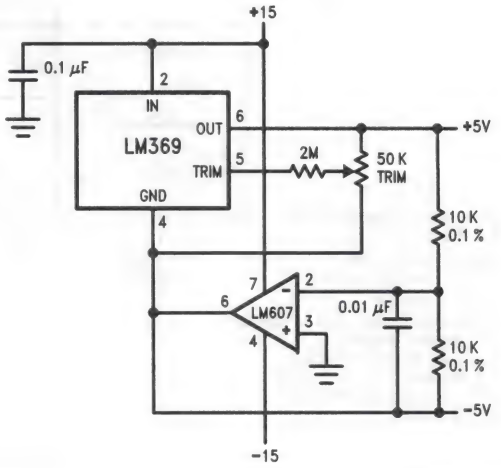
Typical Applications (Continued)

±10V Reference



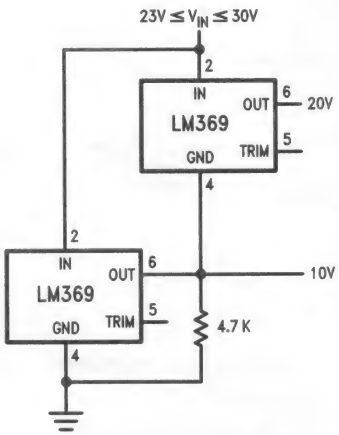
TL/H/9110–7

±5V Reference

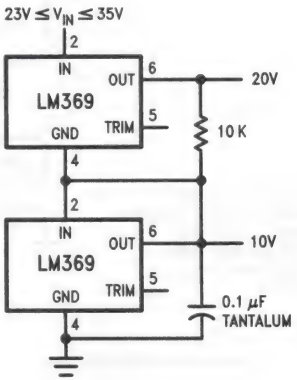


TL/H/9110–8

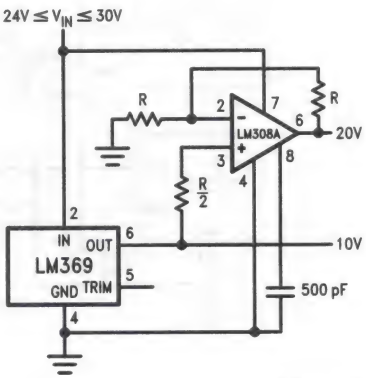
Multiple Output Voltages



TL/H/9110–9



TL/H/9110–10



TL/H/9110–11

R = Thin Film Resistor Network
0.05% Matching and 5 ppm Tracking
(Beckman 694-3-R-10K-A),
(Caddock T-914-10K-100-05)
(Allen Bradley F08B103A)
or similar.

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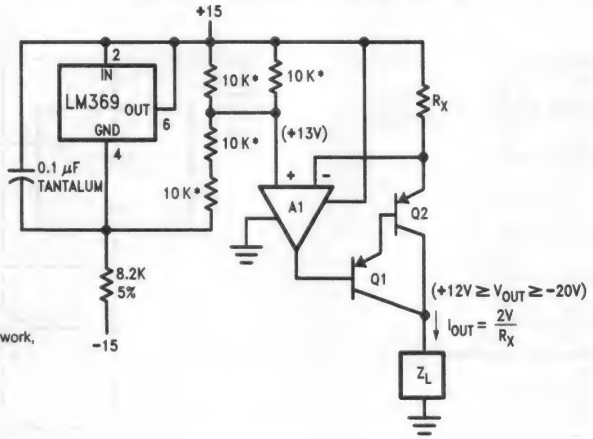
Typical Applications (Continued)

Precision Wide-Range Current Source



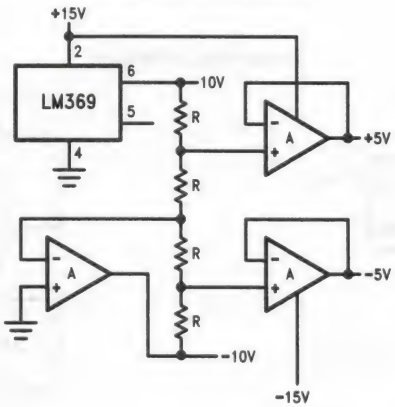
A_1 = LF411A, LM607, LM308A or similar

Q_1, Q_2 = high β PNP, PN4250, 2N3906, or similar

* = Part of Precision Resistor Network, ±0.05% Matching, (Allen Bradley F08B103A) (Caddock T-914-10K-100-05) (Beckman 694-3-R-10K-A) or similar

TL/H/9110–18

±10V, ±5V References

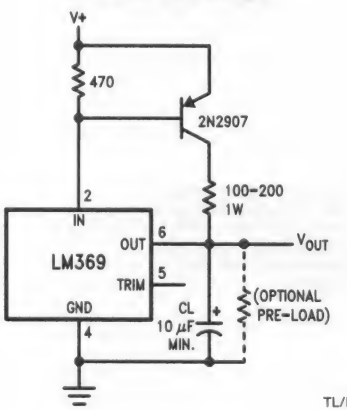


A = ¼ LF444A or ½ LF412A or LM607

R = Thin Film Resistor Network 0.05% Matching and 5 ppm Tracking (Beckman 694-3-R-10K-A), (Caddock T-914-10K-100-05) (Allen Bradley F08B103A) or similar.

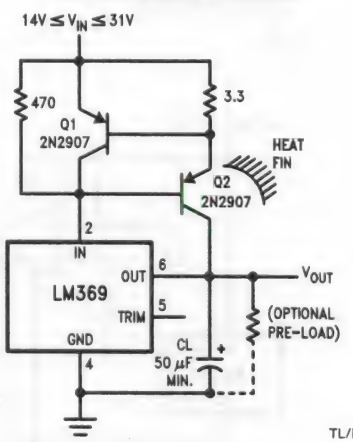
TL/H/9110–12

Reference with Booster



TL/H/9110–13

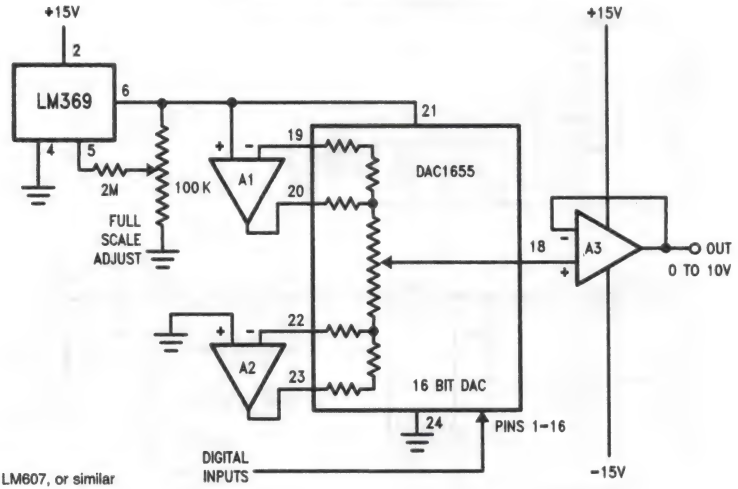
100 mA Boosted Reference



TL/H/9110–14

Typical Applications (Continued)

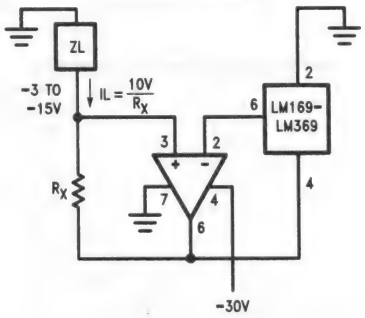
Precision Programmable Supply



A_1, A_2, A_3 = LF411A, LM607, or similar

TL/H/9110-21

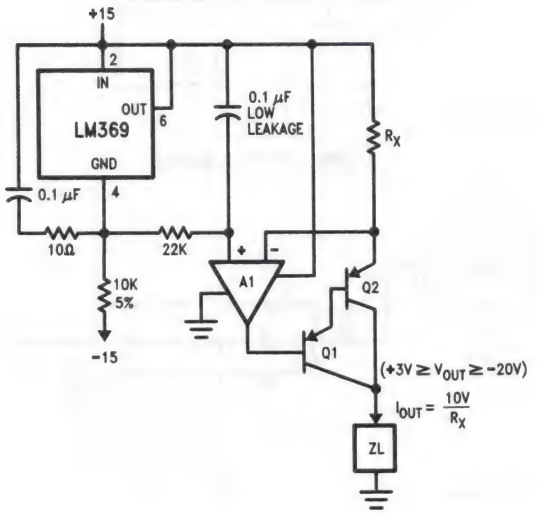
Current Source



$2k \leq Rx \leq 10M$

TL/H/9110-16

Precision Current Source



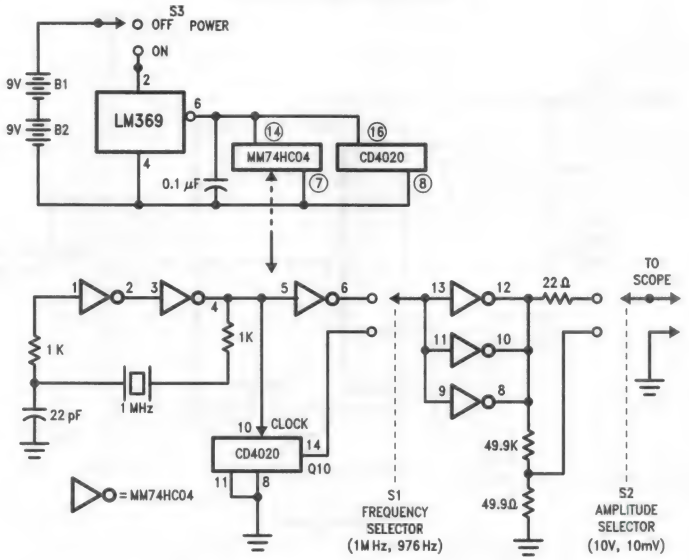
Q_1, Q_2 = high β PNP, PN4250, 2N3906 or similar

A_1 = LM607, LM11, LF411A or similar

TL/H/9110-17

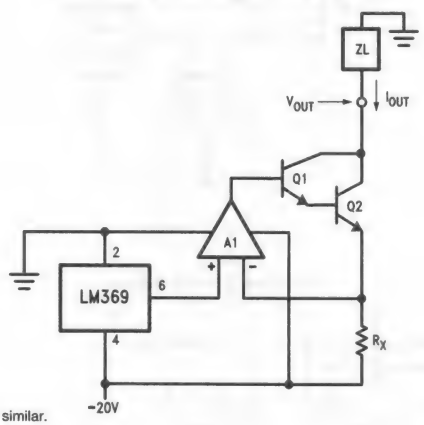
Typical Applications (Continued)

Oscilloscope Calibrator



TL/H/9110–22

Precision Wide-Range Current Sink

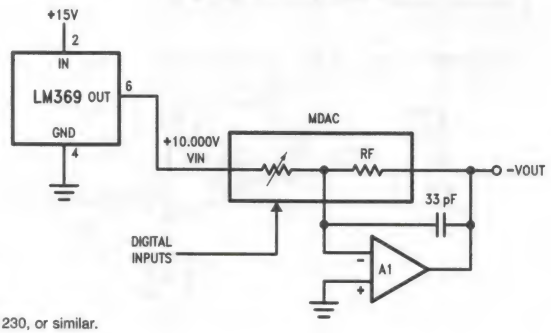


$I_{out} = \frac{10V}{Rx}$

A_1 = LM11, LM607 or similar.
$(V3 + 2V) \leq V_{out} \leq +20V$.
Q1, Q2 = high Beta NPN, 2N3707, 2N3904 or similar.

TL/H/9110–19

Digitally Variable Supply

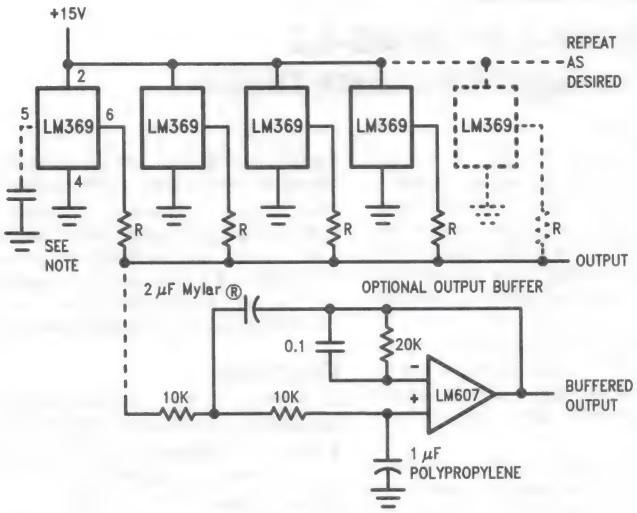


$V_{out} = -10V \times$ (Digitally Set Gain).
A_1 = LM11A, LM607, or similar.
MDAC = DAC1220, DAC1208, DAC1230, or similar.

TL/H/9110–20

Typical Applications (Continued)

Ultra-Low-Noise Statistical Reference



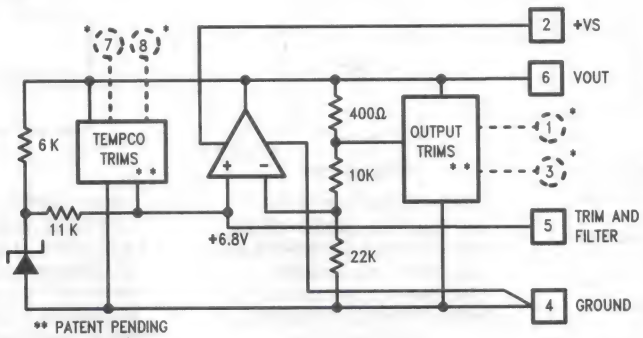
TL/H/9110–23

$200\Omega \leq R \leq 1k$

When N pieces of LM369 are used, the V_{out} noise is decreased by a factor of $\frac{1}{\sqrt{N}}$

If the output buffer is not used, for lowest noise add 0.1 μF Mylar® from ground to pin 5 of each LM369.

LM169 Block Diagram



TL/H/9110–15

*Do not connect; internal connection for factory trim.

National Semiconductor

LM185-1.2/LM285-1.2/LM385-1.2 Micropower Voltage Reference Diode

General Description

The LM185-1.2/LM285-1.2/LM385-1.2 are micropower 2-terminal band-gap voltage regulator diodes. Operating over a 10 μA to 20 mA current range, they feature exceptionally low dynamic impedance and good temperature stability. On-chip trimming is used to provide tight voltage tolerance. Since the LM185-1.2 band-gap reference uses only transistors and resistors, low noise and good long term stability result.

Careful design of the LM185-1.2 has made the device exceptionally tolerant of capacitive loading, making it easy to use in almost any reference application. The wide dynamic operating range allows its use with widely varying supplies with excellent regulation.

The extremely low power drain of the LM185-1.2 makes it useful for micropower circuitry. This voltage reference can be used to make portable meters, regulators or general purpose analog circuitry with battery life approaching shelf life. Further, the wide operating current allows it to replace older references with a tighter tolerance part.

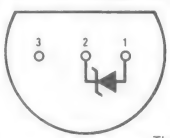
The LM185-1.2 is rated for operation over a −55°C to 125°C temperature range while the LM285-1.2 is rated −40°C to 85°C and the LM385-1.2 0°C to 70°C. The LM185-1.2/LM285-1.2 are available in a hermetic TO-46 package and the LM285-1.2/LM385-1.2 are also available in a low-cost TO-92 molded package, as well as S.O.

Features

- ±4 mV (±0.3%) max. initial tolerance (A grade)
- Operating current of 10 μA to 20 mA
- 0.6Ω max dynamic impedance (A grade)
- Low temperature coefficient
- Low voltage reference—1.235V
- 2.5V device and adjustable device also available — LM185-2.5 series and LM185 series, respectively

Connection Diagrams

TO-92
Plastic Package (Z)

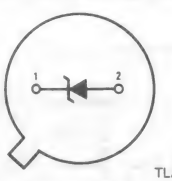


TL/H/5518–10

Bottom View

Order Number LM285Z-1.2, LM285AZ-1.2, LM285AXZ-1.2, LM285AYZ-1.2, LM285BXZ-1.2, LM285BYZ-1.2, LM385Z-1.2, LM385AZ-1.2, LM385AXZ-1.2, LM385AYZ-1.2, LM385BZ-1.2, LM385BXZ-1.2 or LM385BYZ-1.2
See NS Package Number Z03A

TO-46
Metal Can Package (H)

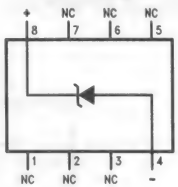


TL/H/5518–6

Bottom View

Order Number LM185H-1.2, LM185AH-1.2, LM185AXH-1.2, LM185AYH-1.2, LM185BXH-1.2, LM185BYH-1.2, LM285H-1.2, LM285AH-1.2, LM285AXH-1.2, LM285AYH-1.2, LM285BXH-1.2 or LM285BYH-1.2
See NS Package Number H02A

SO Package

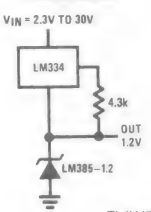


TL/H/5518–9

Order Number LM285M-1.2, LM285AM-1.2, LM285AXM-1.2, LM285AYM-1.2, LM285BXM-1.2, LM285BYM-1.2, LM385M-1.2, LM385AM-1.2, LM385AXM-1.2, LM385AYM-1.2, LM385BM-1.2, LM385BXM-1.2 or LM385BYM-1.2
See NS Package Number M08A

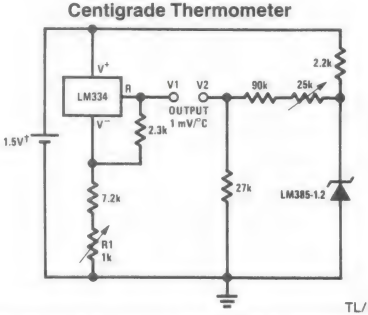
Applications

Wide Input Range Reference



TL/H/5518–8

Centigrade Thermometer



TL/H/5518–1

Calibration

1. Adjust R1 so that V1 = temp at 1 mV/°K
2. Adjust V2 to 273.2 mV

†I_Q for 1.3V to 1.6V battery voltage = 50 μA to 150 μA

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications. (Note 2)

Reverse Current	30 mA
Forward Current	10 mA
Operating Temperature Range (Note 3)	
LM185-1.2	−55°C to +125°C
LM285-1.2	−40°C to +85°C
LM385-1.2	0°C to 70°C
Storage Temperature	−55°C to +150°C
Soldering Information	
TO-92 package: 10 sec.	260°C
TO-46 package: 10 sec.	300°C
SO package: Vapor phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

Electrical Characteristics (Note 4)

Parameter	Conditions	LM185A-1.2 LM185AX-1.2 LM185AY-1.2 LM285A-1.2 LM285AX-1.2 LM285AY-1.2			LM385A-1.2 LM385AX-1.2 LM385AY-1.2			Units (Limit)
		Typ	Tested Limit (Note 5)	Design Limit (Note 6)	Typ	Tested Limit (Note 5)	Design Limit (Note 6)	
Reverse Breakdown Voltage	I_R = 100 μA	1.235	1.231 1.239		1.235	1.231 1.239		V(Min) V(Max)
		1.230		**1.220** **1.245**	**1.235**		**1.225** **1.245**	V(Min) V(Max)
Minimum Operating Current		7	8	**10**	7	8	**10**	μA (Max)
Reverse Breakdown Voltage Change with Current	$I_{MIN} \le I_R \le$ 1 mA		1	**1.5**		1	**1.5**	mV (Max)
	1 mA ≤ $I_R \le$ 20 mA		10	**20**		10	**20**	mV (Max)
Reverse Dynamic Impedance	I_R = 100 μA, f = 20 Hz	0.2		0.6 **1.5**	0.2		0.6 **1.5**	Ω (Max)
Wideband Noise (rms)	I_R = 100 μA, 10 Hz ≤ f ≤ 10 kHz	60			60			μV
Long Term Stability	I_R = 100 μA, T = 1000 Hr, T_A = 25°C ±0.1°C	20			20			ppm
Average Temperature Coefficient (Note 7)	$I_{MIN} \le I_R \le$ 20 mA X Suffix Y Suffix All Others		**30** **50**	 **150**		**30** **50**	 **150**	ppm/°C (Max)

Electrical Characteristics (Continued) (Note 4)

Parameter	Conditions	Typ	LM185-1.2 LM185BX-1.2 LM185BY-1.2 LM285-1.2 LM285BX-1.2 LM285BY-1.2		LM385B-1.2 LM385BX-1.2 LM385BY-1.2		LM385-1.2		Units (Limit)
			Tested Limit (Note 5)	Design Limit (Note 6)	Tested Limit (Note 5)	Design Limit (Note 6)	Tested Limit (Note 5)	Design Limit (Note 6)	
Reverse Breakdown Voltage	$T_A = 25°C$, $10\ \mu A \le I_R \le 20$ mA	1.235	1.223 1.247		1.223 1.247		1.205 1.260		V(Min) V(Max)
Minimum Operating Current		8	10	**20**	15	**20**	15	**20**	μA (Max)
Reverse Breakdown Voltage Change with Current	$10\ \mu A \le I_R \le 1$ mA		1	**1.5**	1	**1.5**	1	**1.5**	mV (Max)
	$1\ mA \le I_R \le 20$ mA		10	**20**	20	**25**	20	**25**	mV (Max)
Reverse Dynamic Impedance	$I_R = 100\ \mu A$, f = 20 Hz	1							Ω
Wideband Noise (rms)	$I_R = 100\ \mu A$, $10\ Hz \le f \le 10$ kHz	60							μV
Long Term Stability	$I_R = 100\ \mu A$, T = 1000 Hr, $T_A = 25°C \pm 0.1°C$	20							ppm
Average Temperature Coefficient (Note 7)	$I_R = 100\ \mu A$ X Suffix Y Suffix All Others		**30** **50**	 **150**	**30** **50**	 **150**		 **150**	 ppm/°C ppm/°C ppm/°C (Max)

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but do not guarantee specific performance limits. For guaranteed specifications and test conditions, see the Electrical Characteristics. The guaranteed specifications apply only for the test conditions listed.

Note 2: Refer to RETS185H-1.2 for military specifications.

Note 3: For elevated temperature operation, T_j max is:

LM185 150°C
LM285 125°C
LM385 100°C

Thermal Resistance	TO-92	TO-46	SO-8
θ_{JA} (junction to ambient)	180°C/W (0.4" leads) 170°C/W (0.125" leads)	440°C/W	165°C/W
θ_{JC} (junction to case)	N/A	80°C/W	N/A

Note 4: Parameters identified with **boldface type** apply at temperature extremes. All other numbers apply at $T_A = T_J = 25°C$.

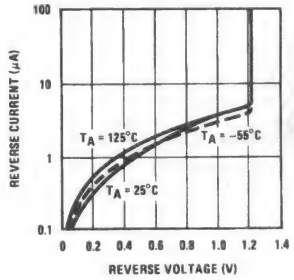
Note 5: Guaranteed and 100% production tested.

Note 6: Guaranteed, but not 100% production tested. These limits are not used to calculate average outgoing quality levels.

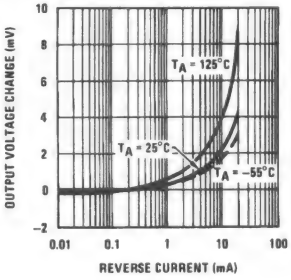
Note 7: The average temperature coefficient is defined as the maximum deviation of reference voltage at all measured temperatures between the operating T_{MAX} and T_{MIN}, divided by $T_{MAX} - T_{MIN}$. The measured temperatures are −55°C, −40°C, 0°C, 25°C, 70°C, 85°C, 125°C.

Typical Performance Characteristics

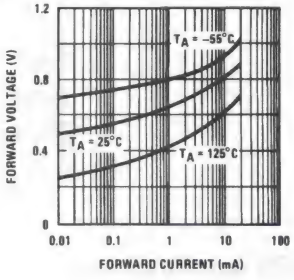
Reverse Characteristics



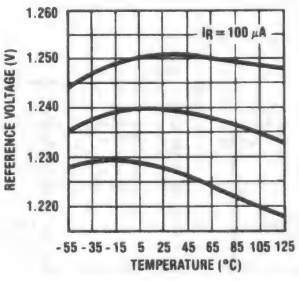
Reverse Characteristics



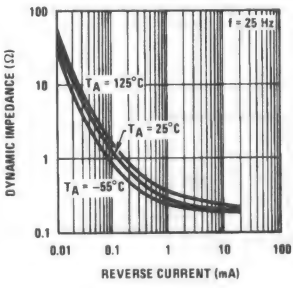
Forward Characteristics



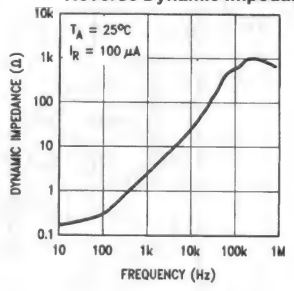
Temperature Drift of 3 Representative Units



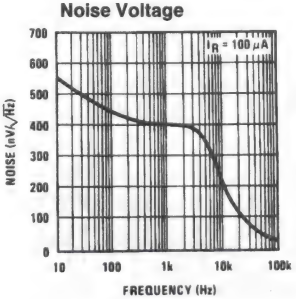
Reverse Dynamic Impedance



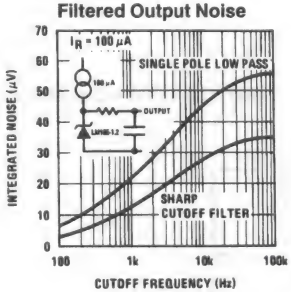
Reverse Dynamic Impedance



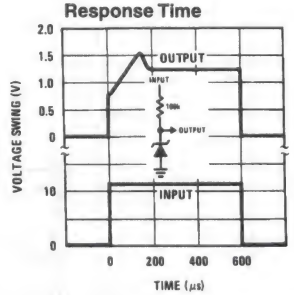
Noise Voltage



Filtered Output Noise



Response Time

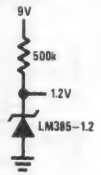


TL/H/5518–3

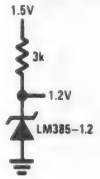
7

Applications (Continued)

Micropower Reference from 9V Battery



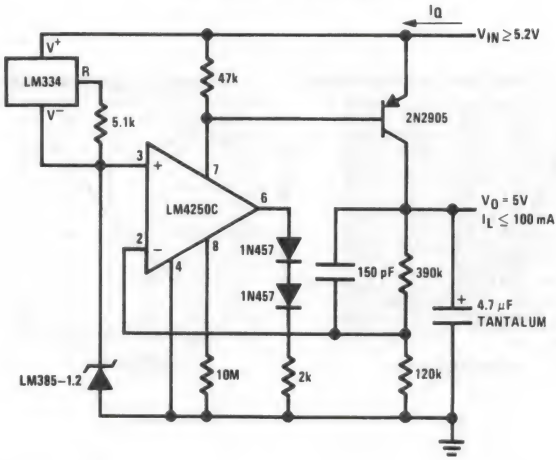
Reference from 1.5V Battery



TL/H/5518–2

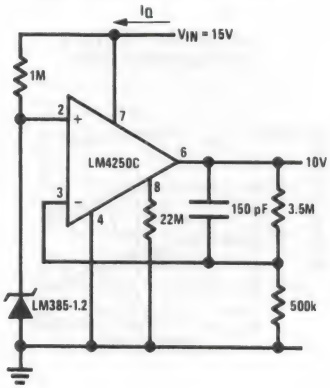
LM385 Applications

Micropower* 5V Regulator



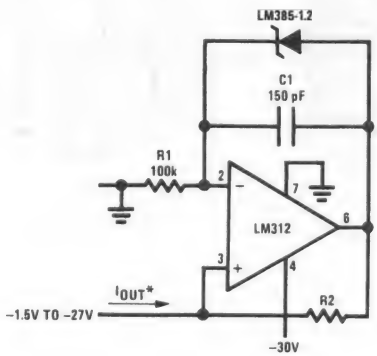
*$I_Q \cong 30$ μA

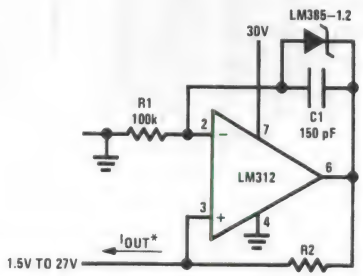
Micropower* 10V Reference



*$I_Q \cong 20$ μA standby current

Precision 1 μA to 1 mA Current Sources





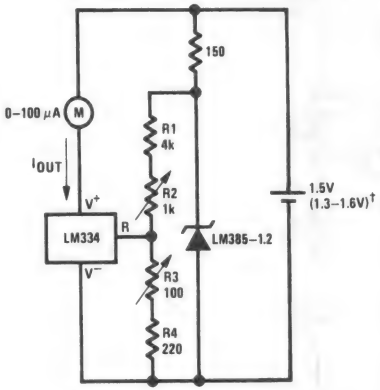
*$I_{OUT} = \frac{1.23V}{R2}$

TL/H/5518–4

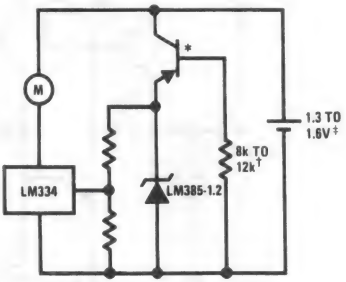
LM385 Applications (Continued)

METER THERMOMETERS

0°C – 100°C Thermometer



Lower Power Thermometer



* 2N3638 or 2N2907 select for inverse $H_{FE} \cong 5$

† Select for operation at 1.3V

‡ $I_Q \cong 600\ \mu A$ to $900\ \mu A$

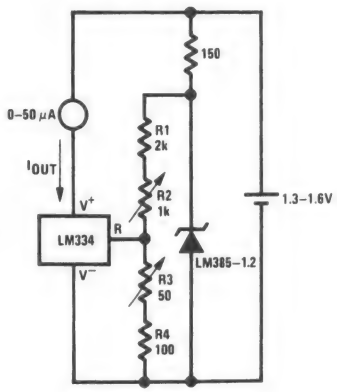
Calibration

1. Short LM385-1.2, adjust R3 for I_{OUT} = temp at 1 μA/°K
2. Remove short, adjust R2 for correct reading in centigrade

†I_Q at 1.3V ≅ 500 μA

I_Q at 1.6V ≅ 2.4 mA

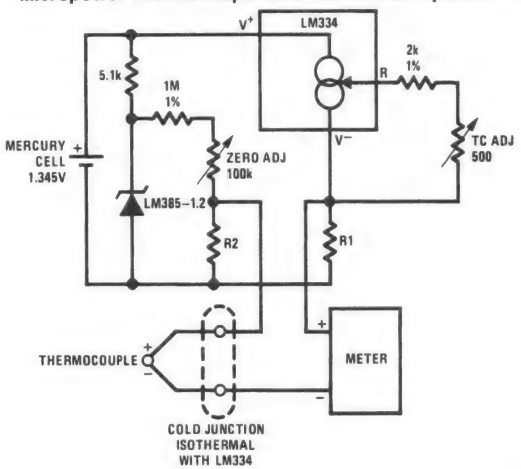
0°F – 50°F Thermometer



Calibration

1. Short LM385-1.2, adjust R3 for I_{OUT} = temp at 1.8 μA/°K
2. Remove short, adjust R2 for correct reading in °F

Micropower Thermocouple Cold Junction Compensator



TL/H/5518–5

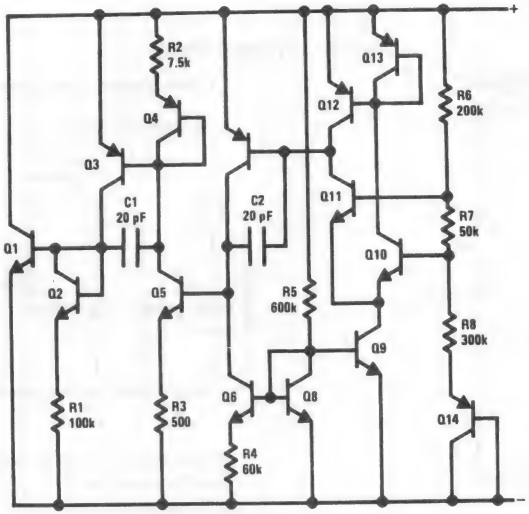
Adjustment Procedure

1. Adjust TC ADJ pot until voltage across R1 equals Kelvin temperature multiplied by the thermocouple Seebeck coefficient.
2. Adjust zero ADJ pot until voltage across R2 equals the thermocouple Seebeck coefficient multiplied by 273.2.

Thermocouple Type	Seebeck Coefficient (μV/°C)	R1 (Ω)	R2 (Ω)	Voltage Across R1 @ 25°C (mV)	Voltage Across R2 (mV)
J	52.3	523	1.24k	15.60	14.32
T	42.8	432	1k	12.77	11.78
K	40.8	412	953Ω	12.17	11.17
S	6.4	63.4	150Ω	1.908	1.766

Typical supply current 50 μA

Schematic Diagram



TL/H/5518–7

National Semiconductor

LM185-2.5/LM285-2.5/LM385-2.5 Micropower Voltage Reference Diode

General Description

The LM185-2.5/LM285-2.5/LM385-2.5 are micropower 2-terminal band-gap voltage regulator diodes. Operating over a 20 μA to 20 mA current range, they feature exceptionally low dynamic impedance and good temperature stability. On-chip trimming is used to provide tight voltage tolerance. Since the LM-185-2.5 band-gap reference uses only transistors and resistors, low noise and good long term stability result.

Careful design of the LM185-2.5 has made the device exceptionally tolerant of capacitive loading, making it easy to use in almost any reference application. The wide dynamic operating range allows its use with widely varying supplies with excellent regulation.

The extremely low power drain of the LM185-2.5 makes it useful for micropower circuitry. This voltage reference can be used to make portable meters, regulators or general purpose analog circuitry with battery life approaching shelf life.

Further, the wide operating current allows it to replace older references with a tighter tolerance part. For applications requiring 1.2V see LM185-1.2.

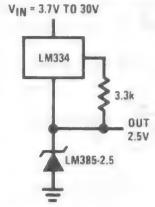
The LM185-2.5 is rated for operation over a −55°C to 125°C temperature range while the LM285-2.5 is rated −40°C to 85°C and the LM385-2.5 0°C to 70°C. The LM185-2.5/LM285-2.5 are available in a hermetic TO-46 package and the LM285-2.5/LM385-2.5 are also available in a low-cost TO-92 molded package, as well as S.O.

Features

- ±20 mV (±0.8%) max. initial tolerance (A grade)
- Operating current of 20 μA to 20 mA
- 0.6Ω dynamic impedance (A grade)
- Low temperature coefficient
- Low voltage reference—2.5V
- 1.2V device and adjustable device also available—LM185-1.2 series and LM185 series, respectively

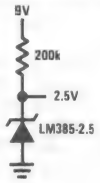
Applications

Wide Input Range Reference



TL/H/5519–12

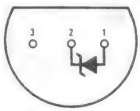
Micropower Reference from 9V Battery



TL/H/5519–2

Connection Diagrams

TO-92
Plastic Package

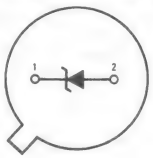


TL/H/5519–8

Bottom View

Order Number LM285Z-2.5, LM285AZ-2.5, LM285AXZ-2.5, LM285AYZ-2.5, LM285BXZ-2.5, LM285BYZ-2.5, LM385Z-2.5, LM385AZ-2.5, LM385AXZ-2.5, LM385AYZ-2.5, LM385BZ-2.5, LM385BXZ-2.5 or LM385BYZ-2.5
See NS Package Number Z03A

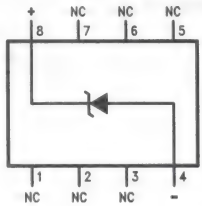
TO-46
Metal Can Package



TL/H/5519–13

Bottom View
Order Number LM185H-2.5, LM185AH-2.5, LM185AXH-2.5, LM185AYH-2.5, LM185BXH-2.5, LM185BYH-2.5, LM285H-2.5, LM285AH-2.5, LM285AXH-2.5, LM285AYH-2.5, LM285BXH-2.5 or LM285BYH-2.5
See NS Package Number H02A

SO Package



TL/H/5519–11

Order Number LM285M-2.5, LM285AM-2.5, LM285AXM-2.5, LM285AYM-2.5, LM285BXM-2.5, LM285BYM-2.5, LM385M-2.5, LM385AM-2.5, LM385AXM-2.5, LM385AYM-2.5, LM385BM-2.5, LM385BXM-2.5 or LM385BYM-2.5
See NS Package Number M08A

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Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications. (Note 2)

Reverse Current	30 mA
Forward Current	10 mA
Operating Temperature Range (Note 3)	
LM185-2.5	−55°C to + 125°C
LM285-2.5	−40°C to + 85°C
LM385-2.5	0°C to 70°C
Storage Temperature	−55°C to + 150°C
Soldering Information	
TO-92 Package (10 sec.)	260°C
TO-46 Package (10 sec.)	300°C
SO Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

Electrical Characteristics (Note 4)

Parameter	Conditions	Typ	LM185A-2.5 LM185AX-2.5 LM185AY-2.5 LM285A-2.5 LM285AX-2.5 LM285AY-2.5		LM385A-2.5 LM385AX-2.5 LM385AY-2.5		Units (Limits)
			Tested Limit (Note 5)	Design Limit (Note 6)	Tested Limit (Note 5)	Design Limit (Note 6)	
Reverse Breakdown Voltage	I_R = 100 μA	2.500	2.480 2.520		2.480 2.520		V(Min) V(Max)
		2.500		**2.460** **2.535**		**2.470** **2.530**	V(Min) V(Max)
Minimum Operating Current		12	18	**20**	18	**20**	μA (Max)
Reverse Breakdown Voltage Change with Current	$I_{MIN} \le I_R \le$ 1mA		1	**1.5**	1	**1.5**	mV (Max)
	1 mA ≤ $I_R \le$ 20 mA		10	**20**	10	**20**	mV (Max)
Reverse Dynamic Impedance	I_R = 100 μA, f = 20 Hz	0.2		0.6 **1.5**		0.6 **1.5**	Ω
Wideband Noise (rms)	I_R = 100 μA 10 Hz ≤ f ≤ 10 kHz	120					μV
Long Term Stability	I_R = 100 μA, T = 1000 Hr, T_A = 25°C ±0.1°C	20					ppm
Average Temperature Coefficient (Note 7)	$I_{MIN} \le I_R \le$ 20 mA X Suffix Y Suffix All Others		**30** **50**	 **150**	**30** **50**	 **150**	ppm/°C (Max)

Electrical Characteristics (Continued) (Note 4)

Parameter	Conditions	Typ	LM185-2.5 LM185BX-2.5 LM185BY-2.5 LM285-2.5 LM285BX-2.5 LM285BY-2.5		LM385B-2.5 LM385BX-2.5 LM385BY-2.5		LM385-2.5		Units (Limit)
			Tested Limit (Note 5)	Design Limit (Note 6)	Tested Limit (Note 5)	Design Limit (Note 6)	Tested Limit (Note 5)	Design Limit (Note 6)	
Reverse Breakdown Voltage	$T_A = 25°C$, $20\ \mu A \le I_R \le 20$ mA	2.5	2.462 2.538		2.462 2.538		2.425 2.575		V(Min) V(Max)
Minimum Operating Current		13	20	**30**	20	**30**	20	**30**	μA (Max)
Reverse Breakdown Voltage Change with Current	$20\ \mu A \le I_R \le 1$ mA		1	**1.5**	2.0	**2.5**	2.0	**2.5**	mV (Max)
	$1\ mA \le I_R \le 20$ mA		10	**20**	20	**25**	20	**25**	mV (Max)
Reverse Dynamic Impedance	$I_R = 100\ \mu A$, $f = 20$ Hz	1							Ω
Wideband Noise (rms)	$I_R = 100\ \mu A$, $10\ Hz \le f \le 10$ kHz	120							μV
Long Term Stability	$I_R = 100\ \mu A$, $T = 1000$ Hr, $T_A = 25°C \pm 0.1°C$	20							ppm
Average Temperature Coefficient (Note 7)	$I_R = 100\ \mu A$ X Suffix Y Suffix All Others		 **30** **50**	 **150**	 **30** **50**	 **150**	 	 **150**	 ppm/°C ppm/°C ppm/°C (Max)

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but do not guarantee specific performance limits. For guaranteed specifications and test conditions, see the Electrical Characteristics. The guaranteed specifications apply only for the test conditions listed.

Note 2: Refer to RETS185H-2.5 for military specifications.

Note 3: For elevated temperature operation, $T_{J\ MAX}$ is:

LM185	150°C
LM285	125°C
LM385	100°C

Thermal Resistance	TO-92	TO-46	SO-8
θ_{ja} (Junction to Ambient)	180°C/W (0.4" Leads) 170°C/W (0.125" Leads)	440°C/W	165°C/W
θ_{jc} (Junction to Case)	N/A	80°C/W	N/A

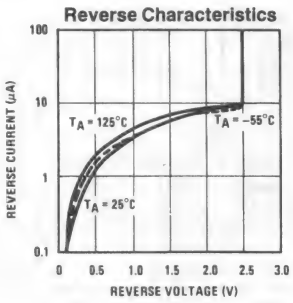
Note 4: Parameters identified with **boldface type** apply at temperature extremes. All other numbers apply at $T_A = T_J = 25°C$.

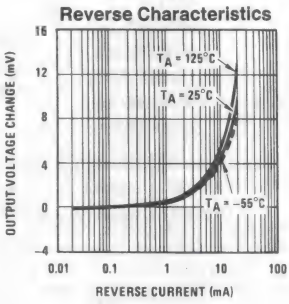
Note 5: Guaranteed and 100% production tested.

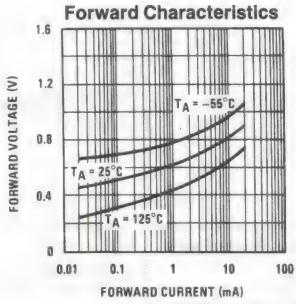
Note 6: Guaranteed, but not 100% production tested. These limits are not used to calculate average outgoing quality levels.

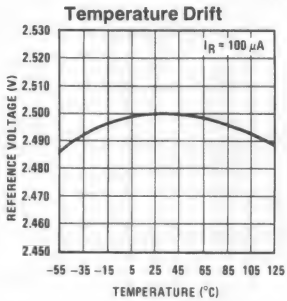
Note 7: The average temperature coefficient is defined as the maximum deviation of reference voltage at all measured temperatures between the operating T_{MAX} and T_{MIN}, divided by $T_{MAX}-T_{MIN}$. The measured temperatures are −55°C, −40°C, 0°C, 25°C, 70°C, 85°C, 125°C.

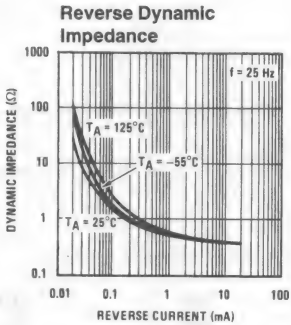
Typical Performance Characteristics

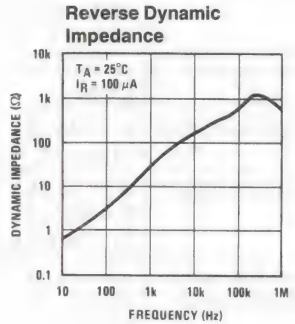


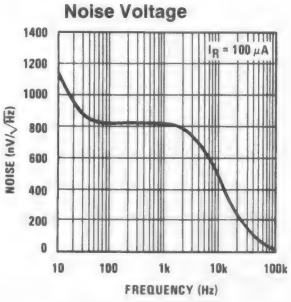


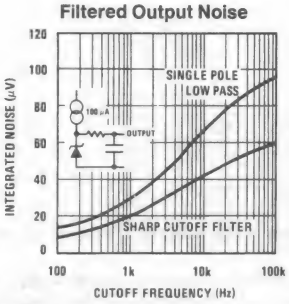


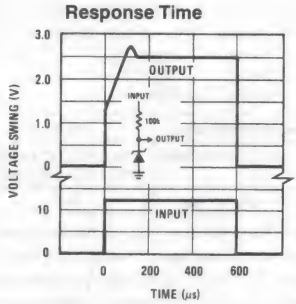








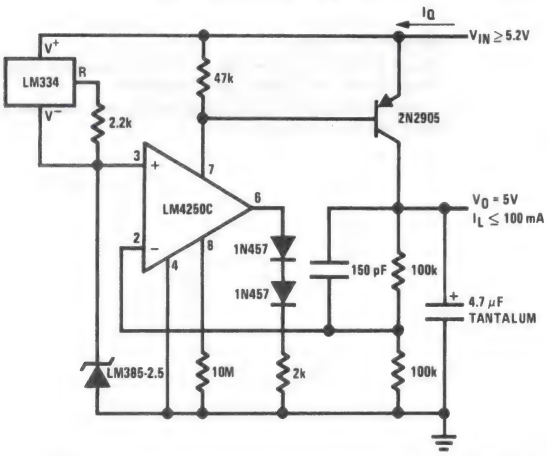




TL/H/5519–3

LM385-2.5 Applications

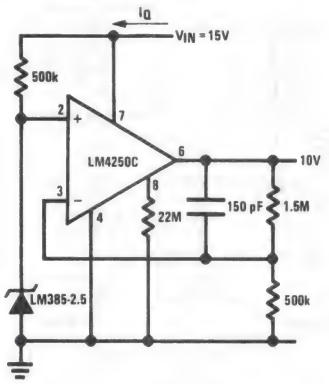
Micropower* 5V Regulator



*$I_Q \cong 40$ μA

TL/H/5519–9

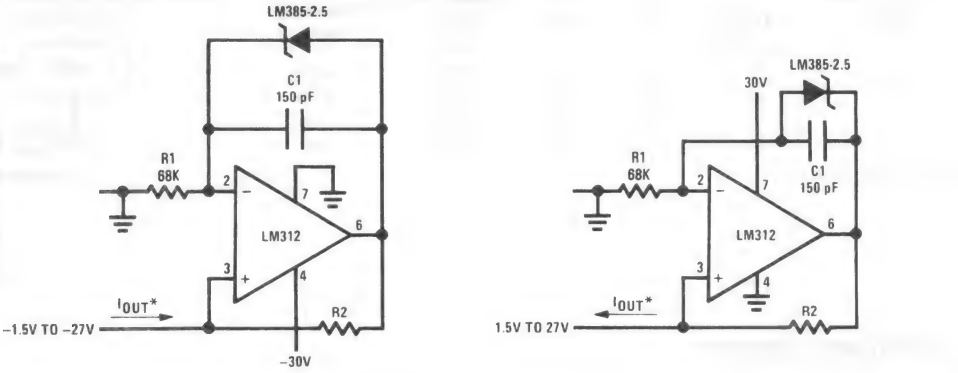
Micropower* 10V Reference



TL/H/5519–10

*$I_Q \cong 30$ μA standby current

Precision 1 μA to 1 mA Current Sources

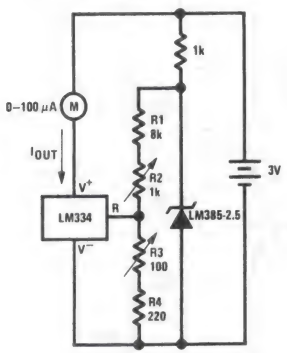


TL/H/5519–4

*$I_{OUT} = \frac{2.5V}{R2}$

METER THERMOMETERS

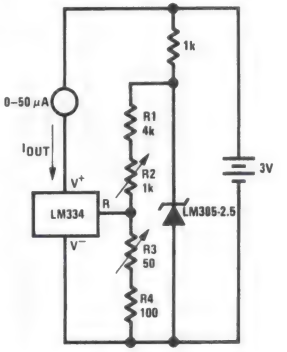
0°C–100°C Thermometer



Calibration

1. Short LM385-2.5, adjust R3 for I_{OUT} = temp at 1μA/°K
2. Remove short, adjust R2 for correct reading in centigrade

0°F–50°F Thermometer



TL/H/5519–5

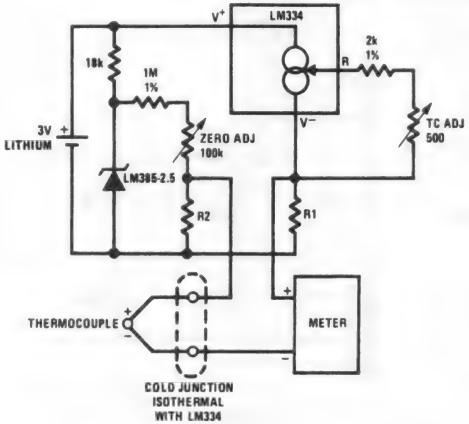
Calibration

1. Short LM385-2.5, adjust R3 for I_{OUT} = temp at 1.8 μA/°K
2. Remove short, adjust R2 for correct reading in °F

7

LM385-2.5 Applications (Continued)

Micropower Thermocouple Cold Junction Compensator



TL/H/5519–6

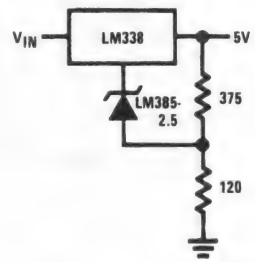
Adjustment Procedure

1. Adjust TC ADJ pot until voltage across R1 equals Kelvin temperature multiplied by the thermocouple Seebeck coefficient.
2. Adjust zero ADJ pot until voltage across R2 equals the thermocouple Seebeck coefficient multiplied by 273.2.

Thermocouple Type	Seebeck Co-efficient (μV/°C)	R1 (Ω)	R2 (Ω)	Voltage Across R1 @25°C (mV)	Voltage Across R2 (mV)
J	52.3	523	1.24k	15.60	14.32
T	42.8	432	1k	12.77	11.78
K	40.8	412	953Ω	12.17	11.17
S	6.4	63.4	150Ω	1.908	1.766

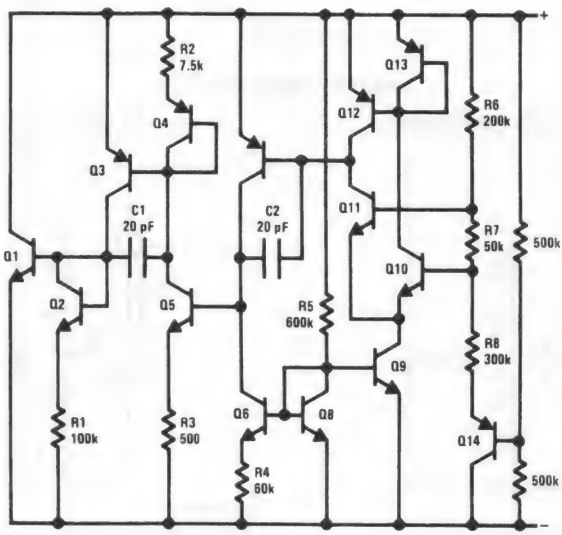
Typical supply current 50 μA

Improving Regulation of Adjustable Regulators



TL/H/5519–7

Schematic Diagram



TL/H/5519–1

National Semiconductor

LM185/LM285/LM385
Adjustable Micropower Voltage References

General Description

The LM185/LM285/LM385 are micropower 3-terminal adjustable band-gap voltage reference diodes. Operating from 1.24 to 5.3V and over a 10 μA to 20 mA current range, they feature exceptionally low dynamic impedance and good temperature stability. On-chip trimming is used to provide tight voltage tolerance. Since the LM185 band-gap reference uses only transistors and resistors, low noise and good long-term stability result.

Careful design of the LM185 has made the device tolerant of capacitive loading, making it easy to use in almost any reference application. The wide dynamic operating range allows its use with widely varying supplies with excellent regulation.

The extremely low power drain of the LM185 makes it useful for micropower circuitry. This voltage reference can be used to make portable meters, regulators or general purpose analog circuitry with battery life approaching shelf life. Further, the wide operating current allows it to replace older references with a tighter tolerance part.

The LM185 is rated for operation over a −55°C to 125°C temperature range, while the LM285 is rated −40°C to 85°C and the LM385 0°C to 70°C. The LM185 is available in a hermetic TO-46 package and the LM285/LM385 are available in a low-cost TO-92 molded package, as well as S.O.

Features

- Adjustable from 1.24V to 5.30V
- Operating current of 10 μA to 20 mA
- 1% and 2% initial tolerance
- 1 Ω dynamic impedance
- Low temperature coefficient

Connection Diagrams

TO-92
Plastic Package

TL/H/5250–9

Bottom View

Order Number LM285BXZ, LM285BYZ, LM285Z, LM385BXZ, LM385BYZ or LM385Z
See NS Package Number Z03A

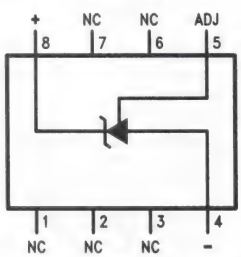
TO-46
Metal Can Package

TL/H/5250–1

Bottom View

Order Number LM185BH, LM185BXH or LM185BYH
See NS Package Number H03A

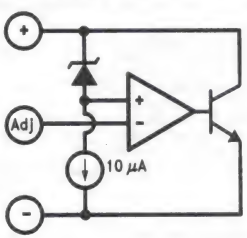
SO Package



TL/H/5250–10

Order Number LM285M or LM385M
See NS Package Number M08A

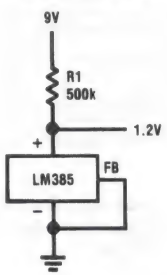
Block Diagram



TL/H/5250–13

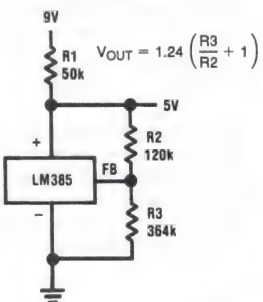
Typical Applications

1.2V Reference



TL/H/5250–14

5.0V Reference



TL/H/5250–2

7

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications. (Note 2)

Reverse Current	30 mA
Forward Current	10 mA
Operating Temperature Range (Note 3)	
LM185 Series	−55°C to 125°C
LM285 Series	−40°C to 85°C
LM385 Series	0°C to 70°C
Storage Temperature	−55°C to 150°C

Soldering Information

TO-92 Package (10 sec.)	260°C
TO-46 Package (10 sec.)	300°C
SO Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

See An-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

Electrical Characteristics (Note 4)

Parameter	Conditions	LM185, LM285					LM385					Units (Limit)
		Typ	LM185BX, LM185BY LM185B, LM285BX, LM285BY		LM285		Typ	LM385BX, LM385BY		LM385		
			Tested Limit (Note 5)	Design Limit (Note 6)	Tested Limit (Note 5)	Design Limit (Note 6)		Tested Limit (Note 5)	Design Limit (Note 6)	Tested Limit (Note 5)	Design Limit (Note 6)	
Reference Voltage	I_R = 100 μA	1.240	1.252 **1.255** 1.228 **1.215**		1.265 1.215	**1.270** **1.205**	1.240	1.252 1.228	**1.255** **1.215**	1.265 1.215	**1.270** **1.205**	V (max) V (min)
Reference Voltage Change with Current	$I_{MIN} < I_R < 1$ mA 1 mA < I_R < 20 mA	0.2 4	1 10	**1.5** **20**	1 10	**1.5** **20**	0.2 5	1 15	**1.5** **25**	1 15	**1.5** **25**	mV (max) mV (max)
Dynamic Output Impedance	I_R = 100 μA, f = 100 Hz I_{AC} = 0.1 I_R $V_{OUT} = V_{REF}$ V_{OUT} = 5.3V	0.3 0.7					0.4 1					Ω
Reference Voltage Change with Output Voltage	I_R = 100 μA	1	3	**6**	3	**6**	2	5	**10**	5	**10**	mV (max)
Feedback Current		13	20	**25**	20	**25**	16	30	**35**	30	**35**	nA (max)
Minimum Operating Current (see curve)	$V_{OUT} = V_{REF}$ V_{OUT} = 5.3V	6 30	9 45	**10** **50**	9 45	**10** **50**	7 35	11 55	**13** **60**	11 55	**13** **60**	μA (max) μA (max)
Output Wideband Noise	I_R = 100 μA, 10 Hz < f < 10 kHz $V_{OUT} = V_{REF}$ V_{OUT} = 5.3V	50 170					50 170					μV_{rms}
Average Temperature Coefficient (Note 7)	I_R = 100 μA X Suffix Y Suffix All Others		**30** **50**	 **150**	**30** **50**	 **150**		**30** **50**	 **150**	**30** **50**	 **150**	ppm/°C (max)
Long Term Stability	I_R = 100 μA, T = 1000 Hr, T_A = 25°C ± 0.1°C	20					20					ppm

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but do not guarantee specific performance limits. For guaranteed specifications and test conditions, see the Electrical Characteristics. The guaranteed specifications apply only for the test conditions listed.

Note 2: Refer to RETS185H for military specifications.

Note 3: For elevated temperature operation, T_j max is:

LM185	150°C
LM285	125°C
LM385	100°C

Thermal Resistance	TO-92	TO-46	SO-8
θ_{ja} (Junction to Ambient)	180°C/W (0.4" leads) 170°C/W (0.125" leads)	440°C/W	165°C/W
θ_{jc} (Junction to Case)	N/A	80°C/W	N/A

Note 4: Parameters identified with **boldface type** apply at temperature extremes. All other numbers apply at $T_A = T_J$ = 25°C. Unless otherwise specified, all parameters apply for $V_{REF} < V_{OUT} < 5.3V$.

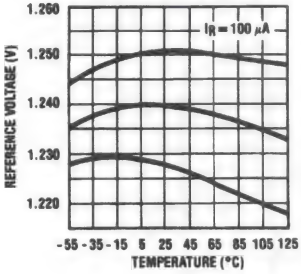
Note 5: Guaranteed and 100% production tested.

Note 6: Guaranteed, but not 100% production tested. These limits are not to be used to calculate average outgoing quality levels.

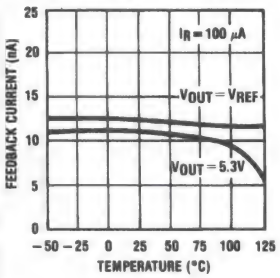
Note 7: The average temperature coefficient is defined as the maximum deviation of reference voltage at all measured temperatures from T_{min} to T_{max}, divided by $T_{max} - T_{min}$. The measured temperatures are −55, −40, 0, 25, 70, 85, 125°C.

Typical Performance Characteristics

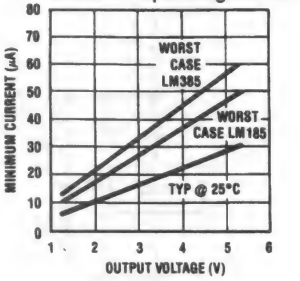
Temperature Drift of 3 Representative Units



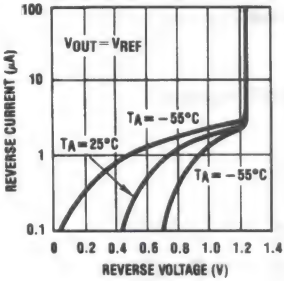
Feedback Current



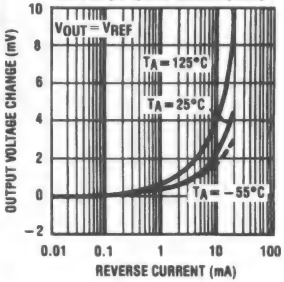
Minimum Operating Current



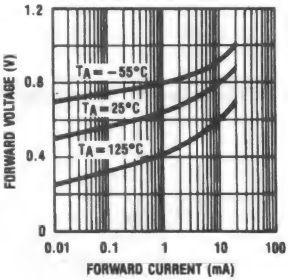
Reverse Characteristics



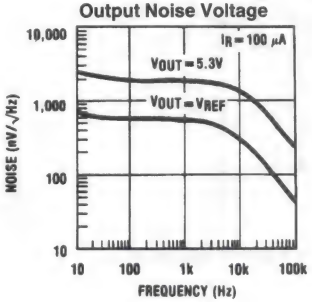
Reverse Characteristics



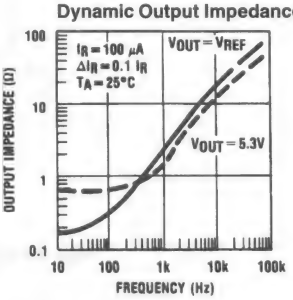
Forward Characteristics



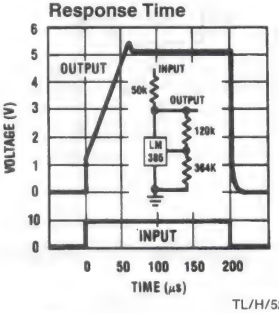
Output Noise Voltage



Dynamic Output Impedance

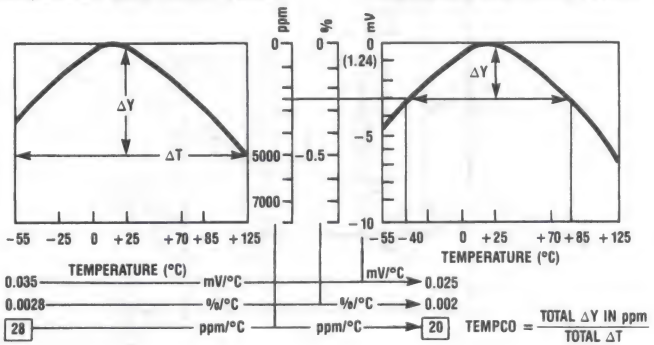


Response Time



TL/H/5250–3

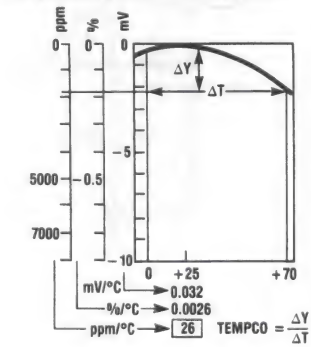
LM185
Temperature Coefficient Typical



LM285
Temperature Coefficient Typical

$$\text{TEMPCO} = \frac{\text{TOTAL } \Delta Y \text{ IN ppm}}{\text{TOTAL } \Delta T}$$

LM385
Temperature Coefficient Typical

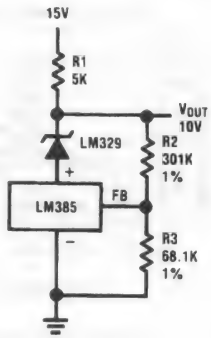


$$\text{TEMPCO} = \frac{\Delta Y}{\Delta T}$$

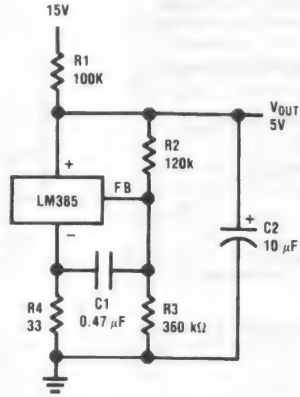
TL/H/5250–4

Typical Applications (Continued)

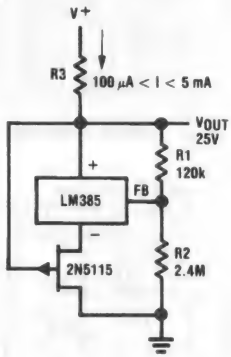
Precision 10V Reference



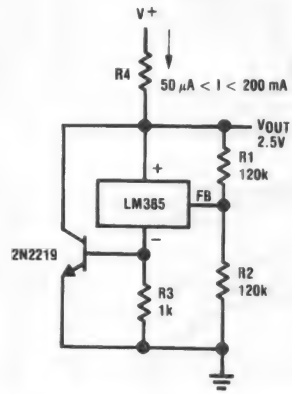
Low AC Noise Reference



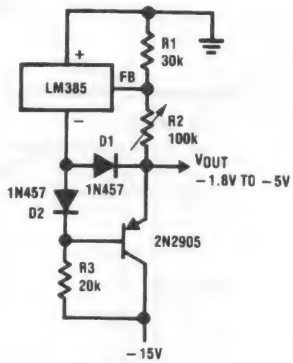
25V Low Current Shunt Regulator



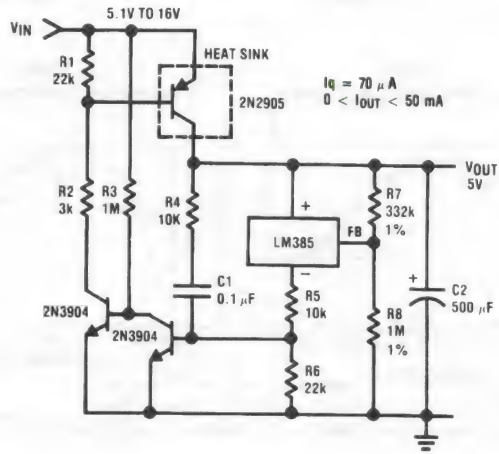
200 mA Shunt Regulator



Series-Shunt 20 mA Regulator



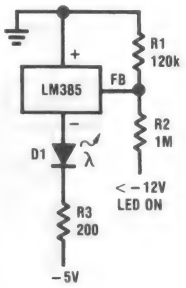
High Efficiency Low Power Regulator



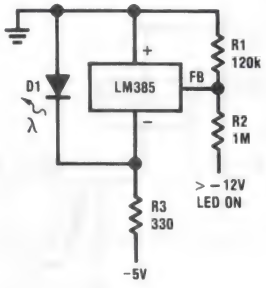
TL/H/5250–5

Typical Applications (Continued)

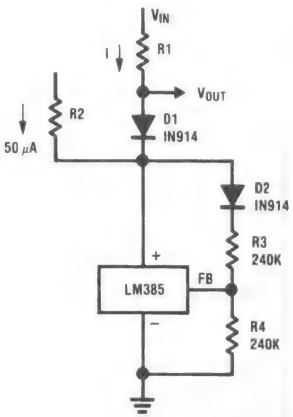
Voltage Level Detector



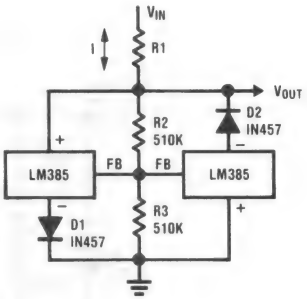
Voltage Level Detector



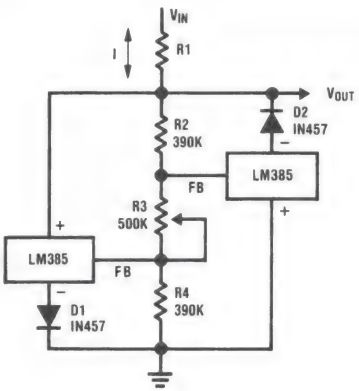
Fast Positive Clamp
2.4V + ΔV_{D1}



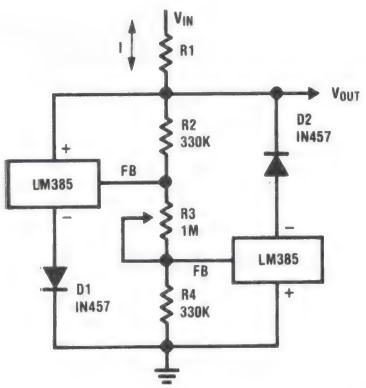
Bidirectional Clamp
±2.4V



Bidirectional Adjustable Clamp
±1.8V to ±2.4V



Bidirectional Adjustable Clamp
±2.4V to ±6V

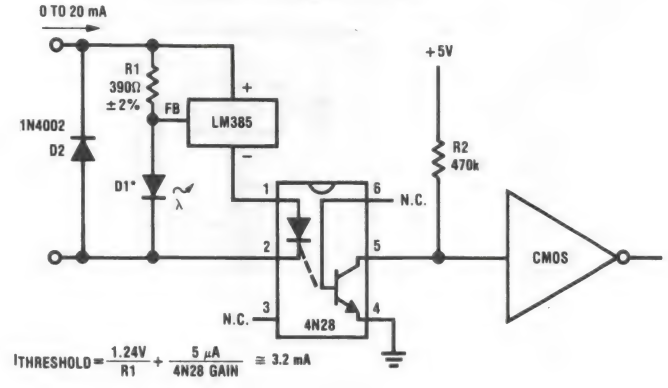


TL/H/5250–6

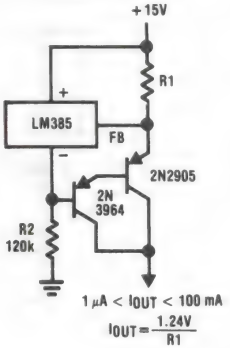
7

Typical Applications (Continued)

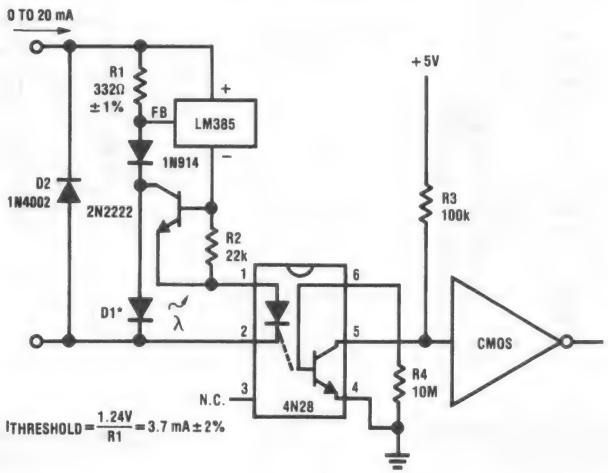
Simple Floating Current Detector



Current Source



Precision Floating Current Detector

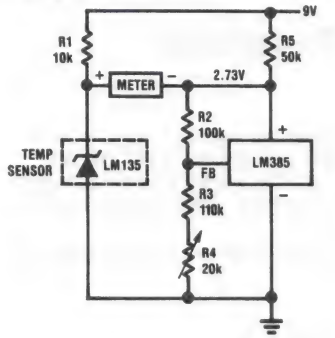


TL/H/5250–7

* D1 can be any LED, $V_F = 1.5V$ to 2.2V at 3 mA. D1 may act as an indicator. D1 will be on if $I_{THRESHOLD}$ falls below the threshold current, except with I = O.

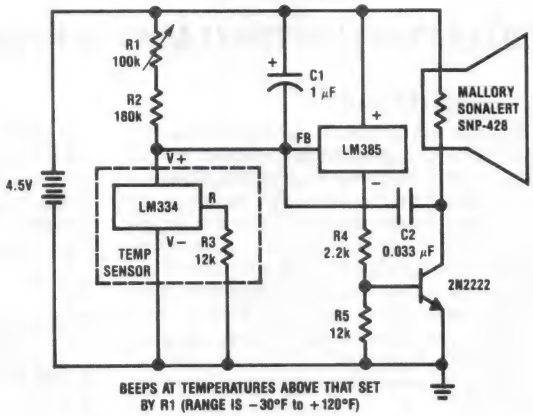
Typical Applications (Continued)

Centigrade Thermometer, 10 mV/°C



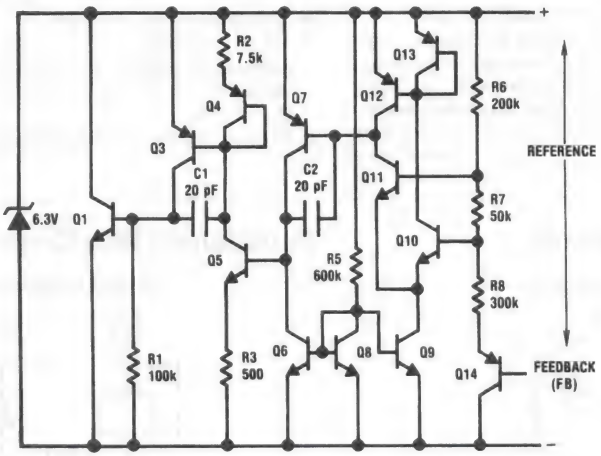
TL/H/5250–11

Freezer Alarm



TL/H/5250–12

Schematic Diagram



TL/H/5250–8

National Semiconductor

LM199/LM299/LM399/LM3999 Precision Reference

General Description

The LM199 series are precision, temperature-stabilized monolithic zeners offering temperature coefficients a factor of ten better than high quality reference zeners. Constructed on a single monolithic chip is a temperature stabilizer circuit and an active reference zener. The active circuitry reduces the dynamic impedance of the zener to about 0.5Ω and allows the zener to operate over 0.5 mA to 10 mA current range with essentially no change in voltage or temperature coefficient. Further, a new subsurface zener structure gives low noise and excellent long term stability compared to ordinary monolithic zeners. The package is supplied with a thermal shield to minimize heater power and improve temperature regulation.

The LM199 series references are exceptionally easy to use and free of the problems that are often experienced with ordinary zeners. There is virtually no hysteresis in reference voltage with temperature cycling. Also, the LM199 is free of voltage shifts due to stress on the leads. Finally, since the unit is temperature stabilized, warm up time is fast.

The LM199 can be used in almost any application in place of ordinary zeners with improved performance. Some ideal applications are analog to digital converters, calibration standards, precision voltage or current sources or precision power supplies. Further in many cases the LM199 can replace references in existing equipment with a minimum of wiring changes.

The LM199 series devices are packaged in a standard hermetic TO-46 package inside a thermal shield. The LM199 is rated for operation from −55°C to +125°C while the LM299 is rated for operation from −25°C to +85°C and the LM399 is rated from 0°C to +70°C.

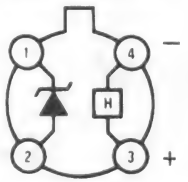
The LM3999 is packaged in a standard TO-92 package and is rated from 0°C to +70°C

Features

- Guaranteed 0.0001%/°C temperature coefficient
- Low dynamic impedance — 0.5Ω
- Initial tolerance on breakdown voltage — 2%
- Sharp breakdown at 400 μA
- Wide operating current — 500 μA to 10 mA
- Wide supply range for temperature stabilizer
- Guaranteed low noise
- Low power for stabilization — 300 mW at 25°C
- Long term stability — 20 ppm
- Proven reliability, low-stress packaging in TO-46 integrated-circuit hermetic package, for low hysteresis after thermal cycling. 33 million hours MTBF at $T_A = +25°C$ ($T_J = +86°C$)
- Certified long term stability available

Connection Diagrams

Metal Can Package

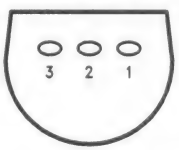


TL/H/5717–14

Top View

LM199/LM299/LM399 (See Table on fourth page)
NS Package Number H04D

Plastic Package TO-92



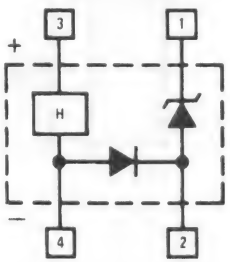
TL/H/5717–10

Bottom View

LM3999 (See Table on fourth page)
NS Package Number Z03A

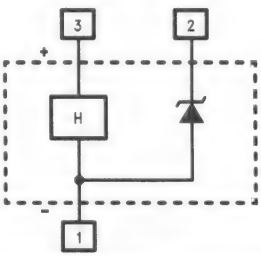
Functional Block Diagrams

LM199/LM299/LM399



TL/H/5717–15

LM3999



TL/H/5717–11

Absolute Maximum Ratings

Specifications for Military/Aerospace products are not contained in this datasheet. Refer to the following Reliability Electrical Test Specifications documents: RETS199X for LM199, RETS199AX for LM199A.

Temperature Stabilizer Voltage
- LM199/LM299/LM399 40V
- LM3999 36V

Reverse Breakdown Current 20 mA

Forward Current
- LM199/LM299/LM399 1 mA
- LM3999 −0.1 mA

Reference to Substrate Voltage $V_{(RS)}$ (Note 1) 40V, −0.1V

Operating Temperature Range
- LM199 −55°C to +125°C
- LM299 −25°C to +85°C
- LM399/LM3999 −0°C to +70°C

Storage Temperature Range −55°C to +150°C

Soldering Information
- TO-92 package (10 sec.) +260°C
- TO-46 package (10 sec.) +300°C

Electrical Characteristics (Note 2)

Parameter	Conditions	LM199/LM299			LM399			Units
		Min	Typ	Max	Min	Typ	Max	
Reverse Breakdown Voltage	$0.5\ mA \le I_R \le 10\ mA$	6.8	6.95	7.1	6.6	6.95	7.3	V
Reverse Breakdown Voltage Change with Current	$0.5\ mA \le I_R \le 10\ mA$		6	9		6	12	mV
Reverse Dynamic Impedance	$I_R = 1\ mA$		0.5	1		0.5	1.5	Ω
Reverse Breakdown Temperature Coefficient	$-55°C \le T_A \le +85°C$ LM199		0.00003	0.0001				%/°C
	$+85°C \le T_A \le +125°C$ LM199		0.0005	0.0015				%/°C
	$-25°C \le T_A \le 85°C$ LM299		0.00003	0.0001				%/°C
	$0°C \le T_A \le +70°C$ LM399					0.00003	0.0002	%/°C
RMS Noise	$10\ Hz \le f \le 10\ kHz$		7	20		7	50	μV
Long Term Stability	Stabilized, $22°C \le T_A \le 28°C$, 1000 Hours, $I_R = 1\ mA \pm 0.1\%$		20			20		ppm
Temperature Stabilizer Supply Current	$T_A = 25°C$, Still Air, $V_S = 30V$ $T_A = -55°C$		8.5 22	14 28		8.5	15	mA
Temperature Stabilizer Supply Voltage		9		40	9		40	V
Warm-Up Time to 0.05%	$V_S = 30V$, $T_A = 25°C$		3			3		sec.
Initial Turn-on Current	$9 \le V_S \le 40$, $T_A = +25°C$, (Note 3)		140	200		140	200	mA

Electrical Characteristics (Note 2)

Parameter	Conditions	LM3999			Units
		Min	Typ	Max	
Reverse Breakdown Voltage	$0.6\ mA \le I_R \le 10\ mA$	6.6	6.95	7.3	V
Reverse Breakdown Voltage Change with Current	$0.6\ mA \le I_R \le 10\ mA$		6	20	mV
Reverse Dynamic Impedance	$I_R = 1\ mA$		0.6	2.2	Ω
Reverse Breakdown Temperature Coefficient	$0°C \le T_A \le 70°C$		0.0002	0.0005	%/°C
RMS Noise	$10\ Hz \le f \le 10\ kHz$		7		μV
Long Term Stability	Stabilized, $22°C \le T_A \le 28°C$, 1000 Hours, $I_R = 1\ mA \pm 0.1\%$		20		ppm
Temperature Stabilizer	$T_A = 25°C$, Still Air, $V_S = 30V$		12	18	mA
Temperature Stabilizer Supply Voltage				36	V
Warm-Up Time to 0.05%	$V_S = 30V$, $T_A = 25°C$		5		sec.
Initial Turn-On Current	$9 \le V_S \le 40$, $T_A = 25°C$		140	200	mA

Electrical Characteristics (Note 2)

Parameter	Conditions	LM199A, LM299A			LM399A			Units
		Min	Typ	Max	Min	Typ	Max	
Reverse Breakdown Voltage	$0.5\ mA \le I_R \le 10\ mA$	6.8	6.95	7.1	6.6	6.95	7.3	V
Reverse Breakdown Voltage Change with Current	$0.5\ mA \le I_R \le 10\ mA$		6	9		6	12	mV
Reverse Dynamic Impedance	$I_R = 1\ mA$		0.5	1		0.5	1.5	Ω
Reverse Breakdown Temperature Coefficient	$-55°C \le T_A \le +85°C$ LM199A		0.00002	0.00005				%/°C
	$+85°C \le T_A \le +125°C$ LM199A		0.0005	0.0010				%/°C
	$-25°C \le T_A \le 85°C$ LM299A		0.00002	0.00005				%/°C
	$0°C \le T_A \le +70°C$ LM399A					0.00003	0.0001	%/°C
RMS Noise	$10\ Hz \le f \le 10\ kHz$		7	20		7	50	μV
Long Term Stability	Stabilized, $22°C \le T_A \le 28°C$, 1000 Hours, $I_R = 1\ mA \pm 0.1\%$		20			20		ppm
Temperature Stabilizer Supply Current	$T_A = 25°C$, Still Air, $V_S = 30V$		8.5	14		8.5	15	mA
	$T_A = -55°C$		22	28				
Temperature Stabilizer Supply Voltage		9		40	9		40	V
Warm-Up Time to 0.05%	$V_S = 30V$, $T_A = 25°C$		3			3		sec.
Initial Turn-on Current	$9 \le V_S \le 40$, $T_A = +25°C$, (Note 3)		140	200		140	200	mA

Electrical Characteristics (Note 2)

Parameter	Conditions	LM199AH-20, LM299AH-20			LM399AH-50			Units
		Min	Typ	Max	Min	Typ	Max	
Reverse Breakdown Voltage	$0.5\ mA \le I_R \le 10\ mA$	6.8	6.95	7.1	6.6	6.95	7.3	V
Reverse Breakdown Voltage Change With Current	$0.5\ mA \le I_R \le 10\ mA$		6	9		6	12	mV
Reverse Dynamic Impedance	$I_R = 1\ mA$		0.5	1		0.5	1.5	Ω
Reverse Breakdown Temperature Coefficient	$-55°C \le T_A \le 85°$ LM199A		0.00002	0.00005				%/°C
	$85°C \le T_A \le 125°C$ LM199A		0.0005	0.0010				%/°C
	$-25°C \le T_A \le 85°C$ LM299A		0.00002	0.00005				%/°C
	$0°C \le T_A \le 70°C$ LM399A					0.00003	0.0001	%/°C
RMS Noise	$10\ Hz \le f \le 10\ kHz$		7	20		7	50	μV
Long Term Stability	Stabilized, $22°C \le T_A \le 28°C$, 1000 Hours, $I_R = 1\ mA \pm 0.1\%$		8	20		9	50	ppm
Temperature Stabilizer Supply Current	$T_A = 25°C$, Still Air, $V_S = 30V$		8.5	14		8.5	15	mA
	$T_A = 55°C$		22	28				
Temperature Stabilizer Supply Voltage		9		40	9		40	V
Warm-Up Time to 0.05%	$V_S = 30V$, $T_A = 25°C$		3			3		s
Initial Turn-on Current	$9 \le V_S \le 40$, $T_A = 25°C$, (Note 3)		140	200		140	200	mA

Note 1: The substrate is electrically connected to the negative terminal of the temperature stabilizer. The voltage that can be applied to either terminal of the reference is 40V more positive or 0.1V more negative than the substrate.

Note 2: These specifications apply for 30V applied to the temperature stabilizer and $-55°C \le T_A \le +125°C$ for the LM199; $-25°C \le T_A \le +85°C$ for the LM299 and $0°C \le T_A \le +70°C$ for the LM399 and LM3999.

Note 3: This initial current can be reduced by adding an appropriate resistor and capacitor to the heater circuit. See the performance characteristic graphs to determine values.

Note 4: Do not wash the LM199 with its polysulfone thermal shield in TCE.

Ordering Information

Initial Tolerance	0°C to +70°C	−25°C to +85°C	−55°C to +125°C	NS Package
2%		LM299AH	LM199AH	H04D
5%	LM399H LM399AH	LM299H	LM199H	H04D
5%	LM3999Z			Z03A
Guaranteed Long Term Stability	LM399AH-50	LM299AH-20	LM199AH-20	H04D

Certified Long Term Drift

The National Semiconductor LM199AH-20, LM299AH-20, and LM399AH-50 are ultra-stable Zener references specially selected from the production runs of LM199AH, LM299AH, LM399AH and tested to confirm a long-term stability of 20, 20, or 50 ppm per 1000 hours, respectively. The devices are measured every 168 hours and the voltage of each device is logged and compared in such a way as to show the deviation from its initial value. Each measurement is taken with a probable-worst-case deviation of ±2 ppm, compared to the Reference Voltage, which is derived from several groups of NBS-traceable references such as LM199AH-20's, 1N827's, and saturated standard cells, so that the deviation of any one group will not cause false indications. Indeed, this comparison process has recently been automated using a specially prepared computer program which is custom-designed to reject noisy data (and require a repeat reading) and to record the average of the best 5 of 7 readings, just as a sagacious standards engineer will reject unbelievable readings.

The typical characteristic for the LM199AH-20 is shown below. This computerized print-out form of each reference's stability is shipped with the unit.

Typical Characteristics

National Semiconductor
Certified Long Term Drift

Hrs	Drift
168	−20
336	−24
504	−36
672	−34
840	−40
1008	−36

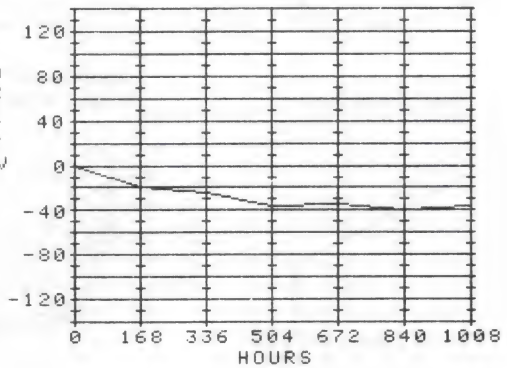
LM199AH-20
Part #6849

Limits

LM199AH-20	140 μV
LM299AH-20	140 μV
LM399AH-20	350 μV

Testing Conditions

Heater Voltage	30V
Zener Current	1 mA
Ambient Temp.	25°C



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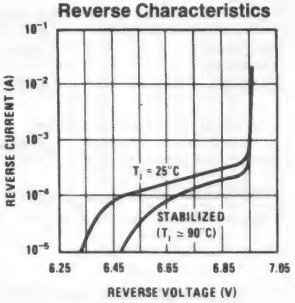
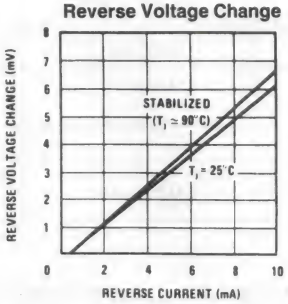
Typical Performance Characteristics

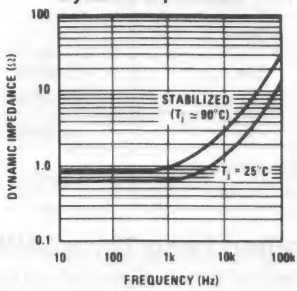
Reverse Characteristics



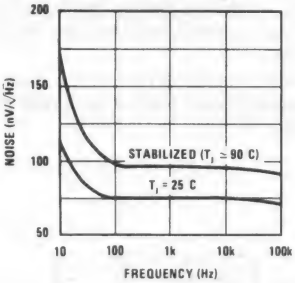
Reverse Voltage Change



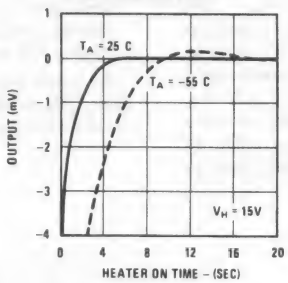
Dynamic Impedance



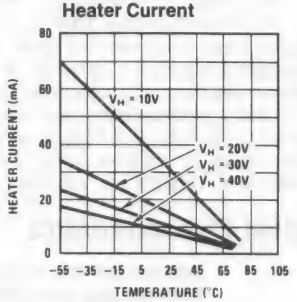
Zener Noise Voltage



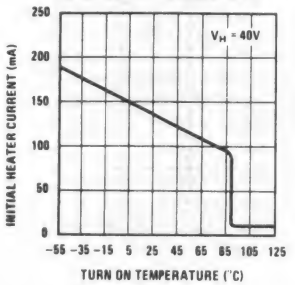
Stabilization Time



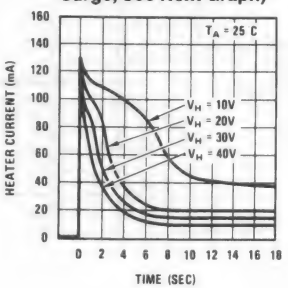
Heater Current



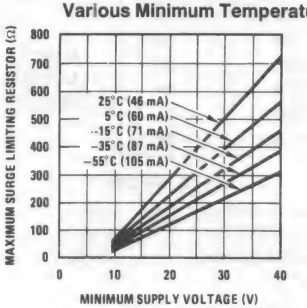
Initial Heater Current



Heater Current (To Limit This Surge, See Next Graph)



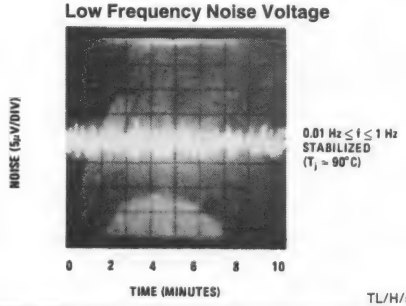
Heater Surge Limit Resistor vs Minimum Supply Voltage at Various Minimum Temperatures



TL/H/5717–2

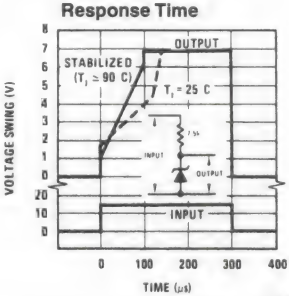
*Heater must be bypassed with a 2 μF or larger tantalum capacitor if resistors are used.

Low Frequency Noise Voltage



TL/H/5717–3

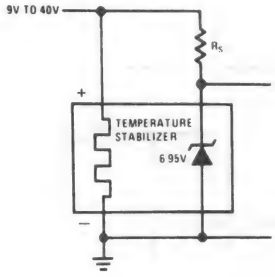
Response Time



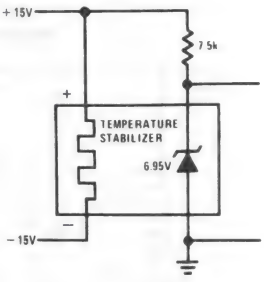
TL/H/5717–7

Typical Applications

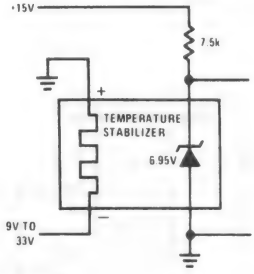
Single Supply Operation



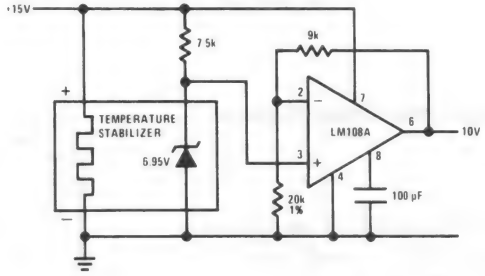
Split Supply Operation



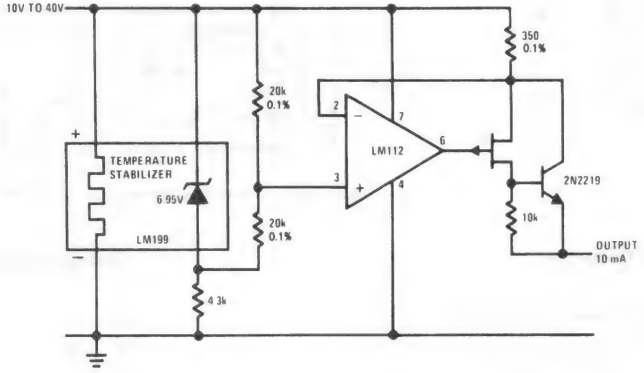
Negative Heater Supply with Positive Reference



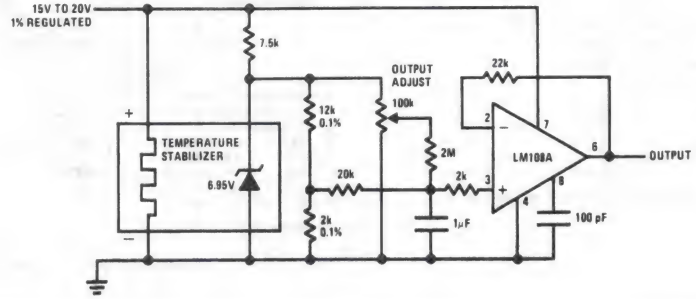
Buffered Reference With Single Supply



Positive Current Source



Standard Cell Replacement



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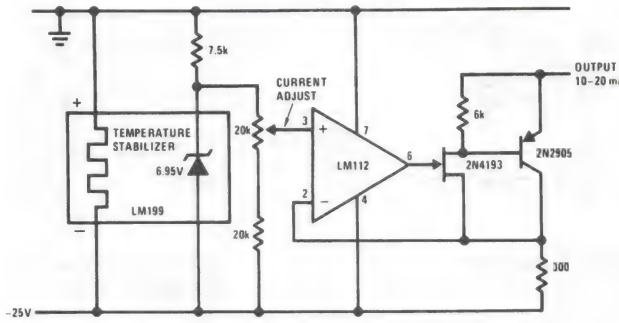
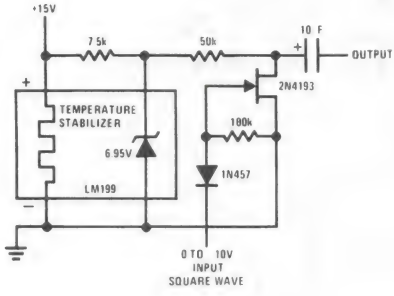
Typical Applications (Continued)

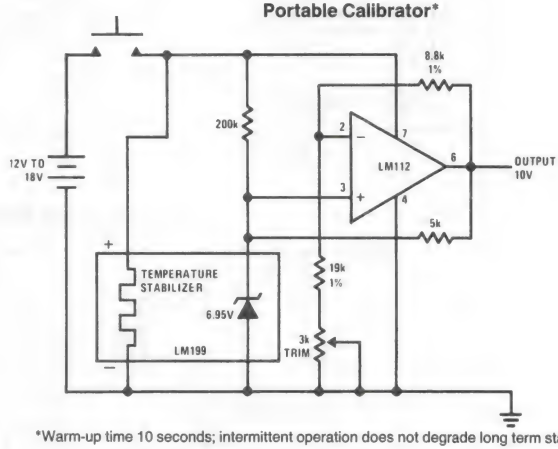
Negative Current Source



Square Wave Voltage Reference

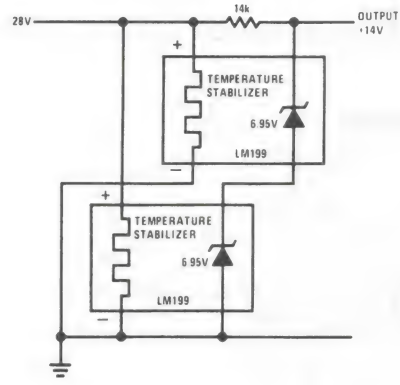


Portable Calibrator*

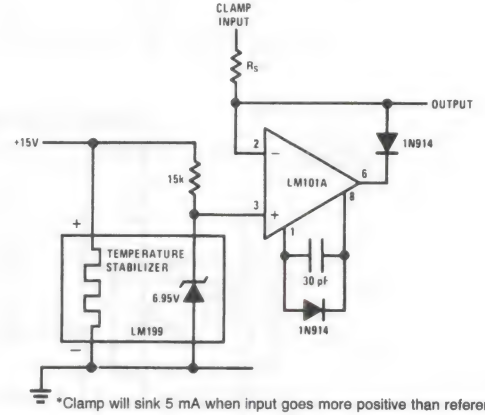


*Warm-up time 10 seconds; intermittent operation does not degrade long term stability.

14V Reference



Precision Clamp*



*Clamp will sink 5 mA when input goes more positive than reference

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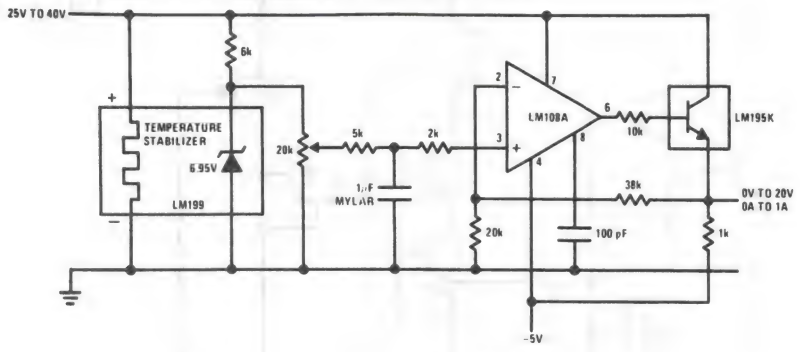
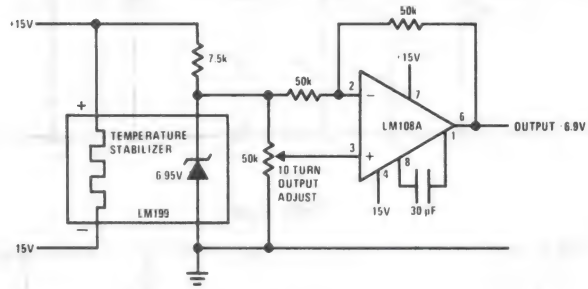
Typical Applications (Continued)

0V to 20V Power Reference



Bipolar Output Reference



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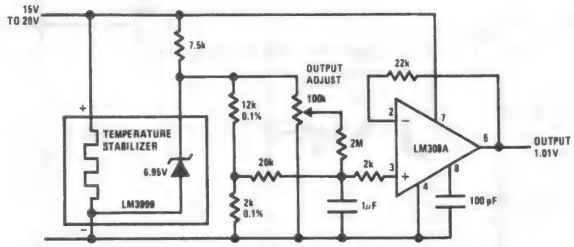
Voltage Reference



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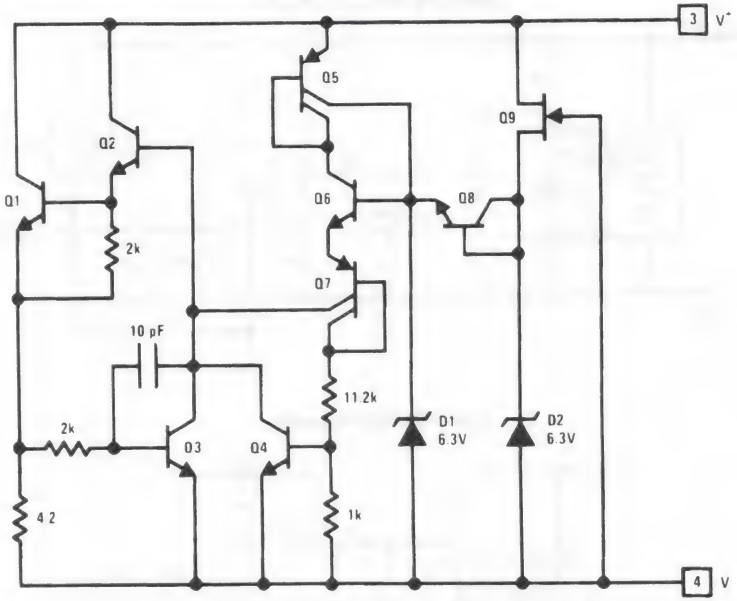
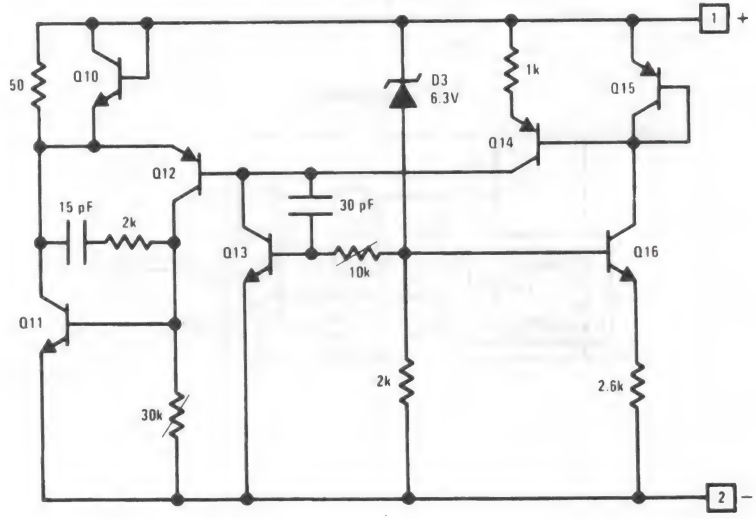
Schematic Diagrams

Temperature Stabilizer



TL/H/5717–01

Reference



TL/H/5717–13

National Semiconductor

PRELIMINARY

LM368-2.5 Precision Voltage Reference

General Description

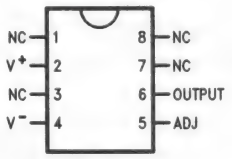
The LM368-2.5 is a precision, monolithic, temperature-compensated voltage reference. The LM368-2.5 makes use of thin-film technology enhanced by the discrete laser trimming of resistors to achieve excellent Temperature coefficient (Tempco) of V_{OUT} (as low as 11 ppm/°C), along with tight initial tolerance, (as low as 0.02%). The trim scheme is such that individual resistors are cut open rather than being trimmed (partially cut), to avoid resistor drift caused by electromigration in the trimmed area. The LM368-2.5 also provides excellent stability vs. changes in input voltage and output current. The output is short circuit proof. A trim pin is made available for fine trimming of V_{OUT} or for obtaining intermediate values without greatly affecting the Tempco of the device.

Features

- 400 μA operating current
- Low output impedance
- Excellent line regulation (.0001%/V typical)
- Single-supply operation
- Externally trimmable
- Low temperature coefficient
- Excellent initial accuracy (0.02% typical)
- Best reference available for low-voltage operation (V_S = 5V, V_{REF} = 2.500V)

Connection Diagrams

Dual-In-Line Package (N)
or S.O. Package (M)

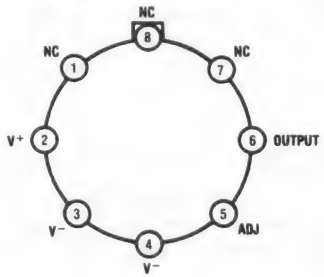


TL/H/8446–15

Top View

Order Number LM368N-2.5
See NS Package Number N08E

Metal Can Package



TL/H/8446–1

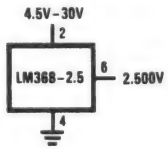
Top View

*case connected to V−

Order Number LM368H-2.5 LM368YH-2.5
See NS Package Number H08C

Typical Applications

Low Voltage Reference



TL/H/8446–2

Absolute Maximum Ratings (Note 7)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Input Voltage	35V
Power Dissipation	600 mW
Storage Temperature Range	−60°C to +150°C
Operating Temperature Range	0°C to +70°C

Soldering Information

DIP (N) Package (10 sec.)	+260°C
TO-5 (H) Package (10 sec.)	+300°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" (Appendix D) for other methods of soldering surface mount devices.

Electrical Characteristics (Note 1)

Parameter	Conditions	LM368-2.5			
		Typical	Tested Limit (Note 2)	Design Limit (Note 3)	Units (Max. unless noted)
V_{OUT} Error: LM368		±0.02	±0.2		%
Line Regulation	$5.0V \le V_{IN} \le 30V$	±0.0001	±0.0005		%/V
Load Regulation (Note 8)	$0\ mA \le I_{SOURCE} \le 10\ mA$	±0.0003	±0.0025		%/mA
Thermal Regulation	T = 20 mS (Note 4)	±0.005	±0.02		%/100 mW
Quiescent Current		350	550		μA
Change of Quiescent Current vs. V_{IN}	$5.0V \le V_{IN} \le 30V$	3	5		μA/V
Temperature Coefficient of V_{OUT} (see graph): LM368Y-2.5	$0°C \le T_A \le 70°C$	±11	±20		ppm/°C
(Note 5) LM368-2.5	$0°C \le T_A \le 70°C$	±15		±30	ppm/°C
Short Circuit Current	$V_{OUT} = 0$	30	70	100	mA
Noise: 0.1–10 Hz		12			uVp-p
100 Hz–10 kHz		420			$nV/\sqrt{Hz}$
V_{OUT} Adjust Range	$0 \le V_{PIN5} \le V_{OUT}$	1.9–5.2		2.2–5.0	V min.

Note 1: Unless otherwise noted, these specifications apply: $T_A = 25°C$, $4.9V \le V_{IN} \le 10.5V$, $0 \le I_{LOAD} \le 0.5\ mA$, $0 \le C_L \le 200\ pF$.

Note 2: Tested Limits are guaranteed and 100% tested in production.

Note 3: Design Limits are guaranteed (but not 100% production tested) over the indicated temperature and supply voltage ranges. These limits are not used to calculate outgoing quality levels.

Note 4: Thermal Regulation is defined as the change in the output Voltage at a time T after a step change in power dissipation of 100 mW.

Note 5: Temperature Coefficient of V_{OUT} is defined as the worst case delta-V_{OUT} measured at Specified Temperatures divided by the total span of the Specified Temperature Range (See graphs). There is no guarantee that the Specified Temperatures are exactly at the minimum or maximum deviation.

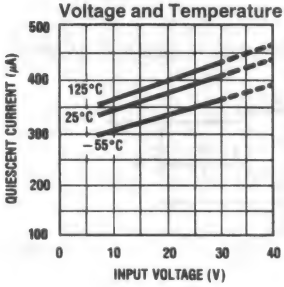
Note 6: In metal can (H), $\theta_{J\text{-}C}$ is 75°C/W and $\theta_{J\text{-}A}$ is 150°C/W. In plastic DIP, $\theta_{J\text{-}A}$ is 160°C/W. In SO-8, $\theta_{J\text{-}A}$ is 180°C/W, in TO-92, $\theta_{J\text{-}A}$ is 160°C/W.

Note 7: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its Rated Operating Conditions (see Note 1 and Conditions).

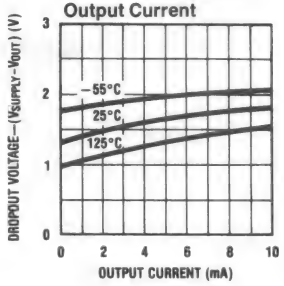
Note 8: Load regulation is measured on the output pin at a point ⅛" below the base of the package. Regulation is measured at constant junction temperature, using pulse testing with a low duty cycle. Changes in output voltage due to heating effects are covered under the specification for thermal regulation.

Typical Performance Characteristics (Note 1)

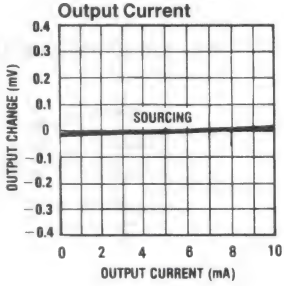
Quiescent Current vs. Input Voltage and Temperature



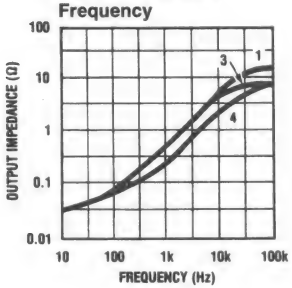
Dropout Voltage vs. Output Current



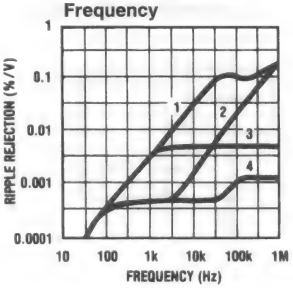
Output Change vs. Output Current



Output Impedance vs. Frequency

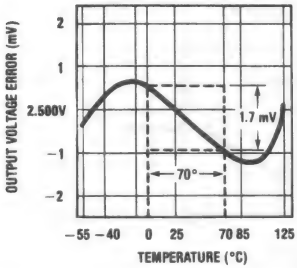


Ripple Rejection vs. Frequency

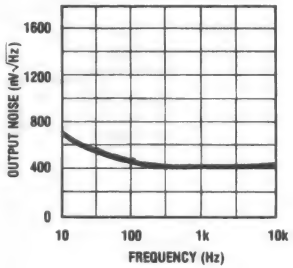


(1) LM368 as is.
(2) with 0.01 μf Mylar, Trim to Gnd.
(3) with 10Ω in series with 10 μf, V_{OUT} to Gnd.
(4) with Both.

Temperature Coefficient: LM368-2.5 (Curve A)



Output Noise vs. Frequency

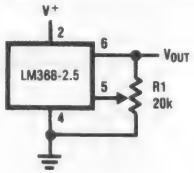


TL/H/8446–3

Typical Temperature Coefficient Calculations:
LM368-2.5 (see Curve A)
T.C. = 1.7 mV/(70° × 2.5V)
= 9.7 ppm/°C

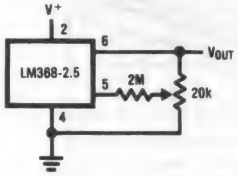
Typical Applications

Wide Range Trimmable Regulator



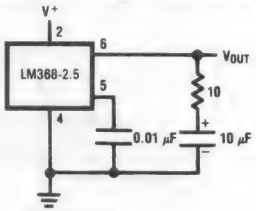
TL/H/8446–5

Narrow Range Trimmable Regulator (±1% min.)



TL/H/8446–6

Improved Noise Performance



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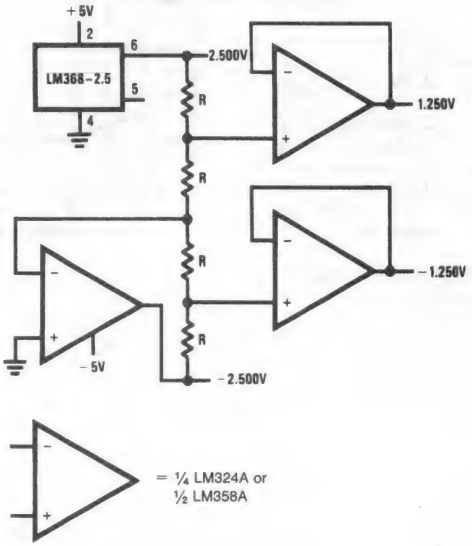
±2.5V, ±1.25V References

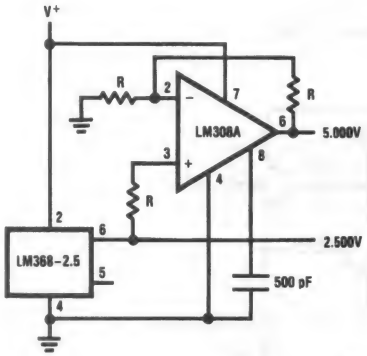


TL/H/8446–8

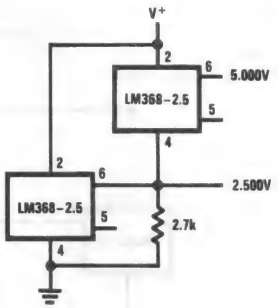
R = Thin Film Resistor Network,
±0.05% Matching and 5 ppm Tracking
(Beckman 694-3-R-10K-A),
(Caddock T-914-10K-100-05)
or similar.

Typical Applications (Continued)

Multiple Output Voltages



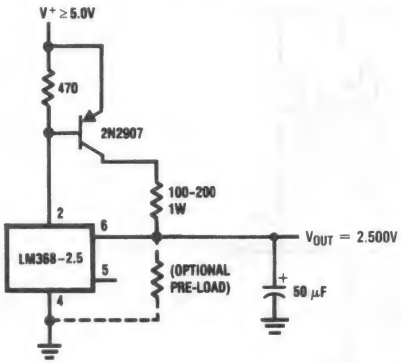
TL/H/8446–10



TL/H/8446–9

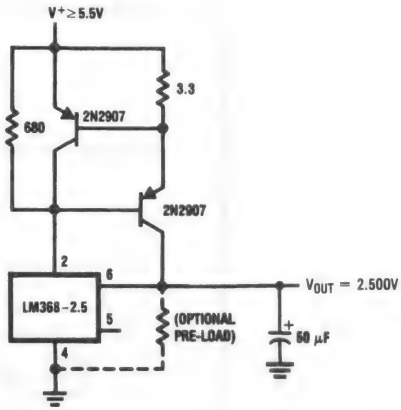
R = Thin Film Resistor Network
0.05% Matching and 5 ppm Tracking
(Beckman 694-3-R-10K-A),
(Caddock T-914-10K-100-05)
or similar.

Reference with Booster



TL/H/8446–11

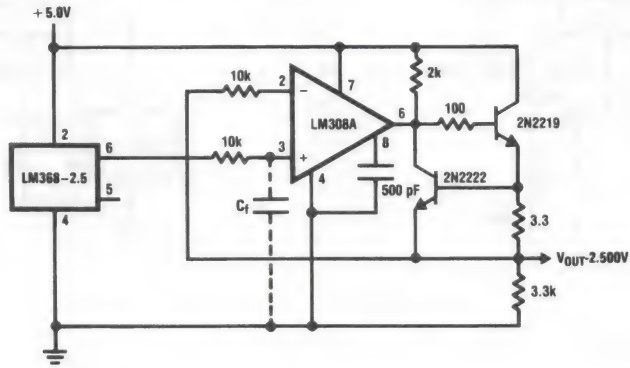
100 mA Boosted Reference



TL/H/8446–12

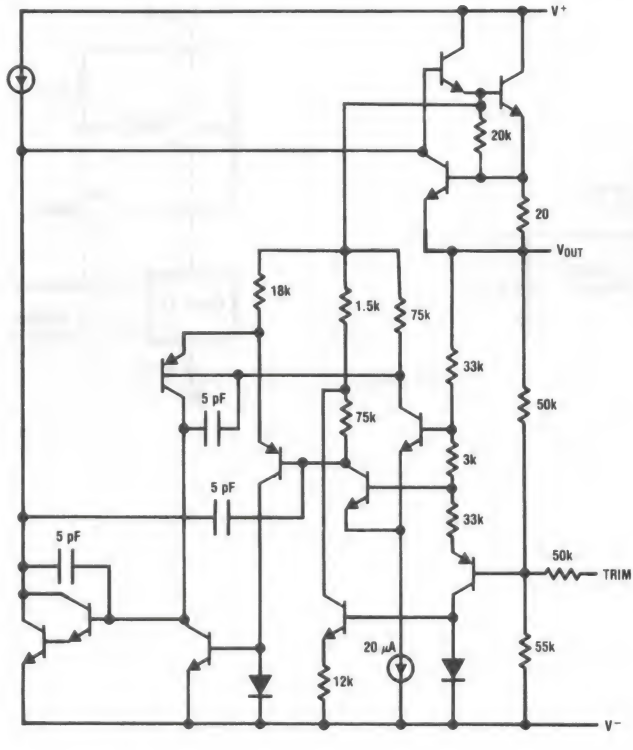
Typical Applications (Continued)

Buffered High-Current Reference with Filter



TL/H/8446–13

Simplified Schematic Diagram



TL/H/8446–14

*Reg. U.S. Pat. Off.

Section 8

Surface Mount

Section 8 Contents

Surface Mount 8-3
AN-450 Small Outline (SO) Package Surface Mounting Methods-Parameters and Their Effect on Product Reliability 8-13

National Semiconductor

Surface Mount

Cost pressures today are forcing many electronics manufacturers to automate their production lines. Surface mount technology plays a key role in this cost-savings trend because:

1. The mounting of devices on the PC board surface eliminates the expense of drilling holes;
2. The use of pick-and-place machines to assemble the PC boards greatly reduces labor costs;
3. The lighter and more compact assembled products resulting from the smaller dimensions of surface mount packages mean lower material costs.

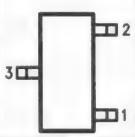
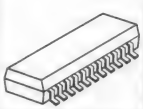
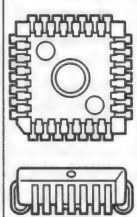
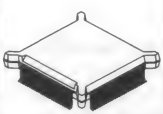
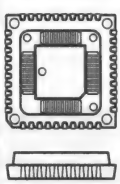
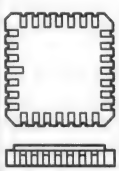
Production processes now permit both surface mount and insertion mount components to be assembled on the same PC board.

SURFACE MOUNT PACKAGING AT NATIONAL

To help our customers take advantage of this new technology, National has developed a line of surface mount packages. Ranging in lead counts from 3 to 360, the package offerings are summarized in Table I.

Lead center spacing keeps shrinking with each new generation of surface mount package. Traditional packages (e.g., DIPs) have a 100 mil lead center spacing. Surface mount packages currently in production (e.g., SOT, SOIC, PCC, LCC, LDCC) have a 50 mil lead center spacing. Surface mount packages in production release (e.g., PQFP) have a 25 mil lead center spacing. Surface mount packages in development (e.g., TAPEPAK®) will have a lead center spacing of only 12–20 mils.

TABLE I. Surface Mount Packages from National

Package Type	Small Outline Transistor (SOT)	Small Outline IC (SOIC)	Plastic Chip Carrier (PCC)	Plastic Quad Flat Pack (PQFP)	TAPEPAK® (TP)	Leadless Chip Carrier (LCC) (LDCC)	Leaded Chip Carrier
							
Package Material	Plastic	Plastic	Plastic	Plastic	Plastic	Ceramic	Ceramic
Lead Bend	Gull Wing	Gull Wing	J-Bend	Gull Wing	Gull Wing	—	Gull Wing
Lead Center Spacing	50 Mils	50 Mils	50 Mils	25 Mils	20, 15, 12 Mils	50 Mils	50 Mils
Tape & Reel Option	Yes	Yes	Yes	tbd	tbd	No	No
Lead Counts	SOT-23 High Profile SOT-23 Low Profile	SO-8(*) SO-14(*) SO-14 Wide(*) SO-16(*) SO-16 Wide(*) SO-20(*) SO-24(*)	PCC-20(*) PCC-28(*) PCC-44(*) PCC-68 PCC-84 PCC-124	PQFP-84 PQFP-100 PQFP-132 PQFP-196(*) PQFP-244	TP-40 (*) TP-68 TP-84 TP-132 TP-172 TP-220 TP-284 TP-360	LCC-18 LCC-20(*) LCC-28 LCC-32 LCC-44 (*) LCC-48 LCC-52 LCC-68 LCC-84 LCC-124	LDCC-44 LDCC-68 LDCC-84 LDCC-124

*In production (or planned) for linear products.

LINEAR PRODUCTS IN SURFACE MOUNT

Linear functions available in surface mount include:

- Op amps
- Comparators
- Regulators
- References
- Data conversion
- Industrial
- Consumer
- Automotive

A complete list of linear part numbers in surface mount is presented in Table III. Refer to the datasheet in the appropriate chapter of this databook for a complete description of the device. In addition, National is continually expanding the list of devices offered in surface mount. If the functions you need do not appear in Table III, contact the sales office or distributor branch nearest you for additional information.

Automated manufacturers can improve their cost savings by using Tape-and-Reel for surface mount devices. Simplified handling results because hundreds-to-thousands of semiconductors are carried on a single Tape-and-Reel pack (see ordering and shipping information—printed later in this section—for a comparison of devices/reel vs. devices/rail for those surface mount package types being used for linear products). With this higher device count per reel (when compared with less than a 100 devices per rail), pick-and-place machines have to be re-loaded less frequently and lower labor costs result.

With Tape-and-Reel, manufacturers save twice—once from using surface mount technology for automated PC board assembly and again from less device handling during shipment and machine set-up.

BOARD CONVERSION

Besides new designs, many manufacturers are converting existing printed circuit board designs to surface mount. The resulting PCB will be smaller, lighter and less expensive to manufacture; but there is one caveat—be careful about the thermal dissipation capability of the surface mount package.

Because the surface mount package is smaller than the traditional dual-in-line package, the surface mount package is not capable of conducting as much heat away as the DIP (i.e., the surface mount package has a higher thermal resistance—see Table II).

The silicon for most National devices can operate up to a 150°C junction temperature (check the datasheet for the rare exception). Like the DIP, the surface mount package can actually withstand an ambient temperature of up to 125°C (although a commercial temperature range device will only be specified for a max ambient temperature of 70°C and an industrial temperature range device will only be specified for a max ambient temperature of 85°C). See AN-336, "Understanding Integrated Circuit Package Power Capabilities", (reprinted in the appendix of each linear databook volume) for more information.

TABLE II: Surface Mount Package Thermal Resistance Range*

Package	Thermal Resistance** (θ_{JA}, °C/W)
SO-8	120–175
SO-14	100–140
SO-14 Wide	70–110
SO-16	90–130
SO-16 Wide	70–100
SO-20	60–90
SO-24	55–85
PCC-20	70–100
PCC-28	60–90
PCC-44	40–60

*Actual thermal resistance for a particular device depends on die size. Refer to the datasheet for the actual θ_{JA} value.

**Test conditions: PCB mount (FR4 material), still air (room temperature), copper traces (150 × 20 × 10 mils).

Given a max junction temperature of 150°C and a maximum allowed ambient temperature, the surface mount device will be able to dissipate less power than the DIP device. This factor must be taken into account for new designs.

For board conversion, the DIP and surface mount devices would have to dissipate the same power. This means the surface mount circuit would have a lower maximum allowable ambient temperature than the DIP circuit. For DIP circuits where the maximum ambient temperature required is substantially lower than the maximum ambient temperature allowed, there may be enough margin for safe operation of the surface mount circuit with its lower maximum allowable ambient temperature. But where the maximum ambient temperature required of the DIP current is close to the maximum allowable ambient temperature, the lower maximum ambient temperature allowed for the surface mount circuit may fall below the maximum ambient temperature required. The circuit designer must be aware of this potential pitfall so that an appropriate work-around can be found to keep the surface mount package from being thermally overstressed in the application.

SURFACE MOUNT LITERATURE

National has published extensive literature on the subject of surface mount packaging. Engineers from packaging, quality, reliability, and surface mount applications have pooled their experience to provide you with practical hands-on knowledge about the construction and use of surface mount packages.

The applications note AN-450 "Surface Mounting Methods and their Effect on Product Reliability" is referenced on each SMD datasheet. In addition, "Wave Soldering of Surface Mount Components" is reprinted in this section for your information.

TABLE III. Linear Surface Mount Current Device Listing

Amplifiers and Comparators

Part Number	Part Number
LF347WM LF351M LF451CM LF353M	LM392M LM393M LM741CM LM1458M LM2901M
LF355M LF356M LF357M LF444CWM	LM2902M LM2903M LM2904M LM2924M LM3403M
LM10CWM LM10CLWM LM308M LM308AM LM310M	LM4250M LM324M LM339M LM365WM LM607CM
LM311M LM318M LM319M LM324M	LMC669BCWM LMC669CCWM
LM339M LM346M LM348M LM358M LM359M	LF441CM

Regulators and References

Part Number	Part Number
LM317LM LF3334M	LM2931M-5.0 LM3524M LM78L05ACM LM78L12ACM LM78L15ACM
LM336M-2.5 LF336BM-2.5 LM336M-5.0 LM336BM-5.0 LM337LM	LM79L05ACM LM79L12ACM LM79L15ACM LP2951ACM LP2951CM
LM385M LM385M-1.2	
LM385BM-1.2 LM385M-2.5 LM385BM-2.5 LM723CM LM2931CM	

Data Acquisition Circuits

Part Number	Part Number
ADC0802LCV ADC0802LCWM ADC0804LCV ADC0804LCWM ADC0808CCV ADC0809CCV	ADC1025BCV ADC1025CCV DAC0800LCM DAC0801LCM DAC0802LCM DAC0806LCM DAC0807LCM
ADC0811BCV ADC0811CCV ADC0819BCV ADC0819CCV ADC0820BCV ADC0820CCV	DAC0808LCM DAC0830LCWM DAC0830LCV DAC0832LCWM DAC0832LCV
ADC0838BCV ADC0838CCV ADC0841BCV ADC0841CCV ADC0848BCV ADC0848CCV ADC1005BCV ADC1005CCV	

Industrial Functions

Part Number	Part Number
AH5012CM LF13331M LF13509M LF13333M LM555CM	LM13600M LM13700M LMC555CM LM567CM MF4CWM-50
LM556CM LM567CM LM1496M LM2917M	MF4CWM-100 MF6CWM-50 MF10CCWM MF6CWM-100 MF5CWM
LM3046M LM3086M LM3146M	

Commercial and Automotive

Part Number	Part Number
LM386M-1 LM592M LM831M LM832M LM833M	LM1837M LM1851M LM1863M LM1865M LM1870M
LM837M LM838M LM1131CM	LM1894M LM1964V
	LM2893M LM3361AM LM1881M

Hybrids

Part Number	Part Number
LH0002E	LH0032E
LH4002E	LH0033E

A FINAL WORD

National is a world leader in the design and manufacture of surface mount components.

Because of design innovations such as perforated copper leadframes, our small outline package is as reliable as our DIP—the laws of physics would have meant that a straight "junior copy" of the DIP would have resulted in an "S.O." package of lower reliability. You benefit from this equivalence of reliability. In addition, our ongoing vigilance at each step of the production process assures that the reliability we designed in stays in so that only devices of the highest quality and reliability are shipped to your factory.

Our surface mount applications lab at our headquarters site in Santa Clara, California continues to research (and publish) methods to make it even easier for you to use surface mount technology. Your problems are our problems.

When you think "Surface Mount"—think "National"!

Ordering and Shipping Information

When you order a surface mount semiconductor, it will be in one of the several available surface mount package types. Specifying the Tape-and-Reel method of shipment means that you will receive your devices in the following quantities per Tape-and-Reel pack: SMD devices can also be supplied in conventional conductive rails.

Package	Package Designator	Max/Rail	Per Reel*
SO-8	M	100	2500
SO-14	M	50	2500
SO-14 Wide	WM	50	1000
SO-16	M	50	2500
SO-16 Wide	WM	50	1000
SO-20	M	40	1000
SO-24	M	30	1000
PCL-20	V	50	1000
PCL-28	V	40	1000
PCL-44	V	25	500
PQFP-196	VF	TBD	—
TP-40	TP	100	TBD
LCC-20	E	50	—
LCC-44	E	25	—

*Incremental ordering quantities. (National Semiconductor reserves the right to provide a smaller quantity of devices per Tape-and-Reel pack to preserve lot or date code integrity. See example below.)

Example: You order 5,000 LM324M ICs shipped in Tape-and-Reel.

- Case 1: All 5,000 devices have the same date code
  - You receive 2 SO-14 (Narrow) Tape-and-Reel packs, each having 2500 LM324M ICs
- Case 2: 3,000 devices have date code A and 2,000 devices have date code B
  - You receive 3 SO-14 (Narrow) Tape-and-Reel packs as follows:

    Pack #1 has 2,500 LM324M ICs with date code A

    Pack #2 has 500 LM324M ICs with date code A

    Pack #3 has 2,000 LM324M ICs with date code B

Short-Form Procurement Specification

TAPE FORMAT

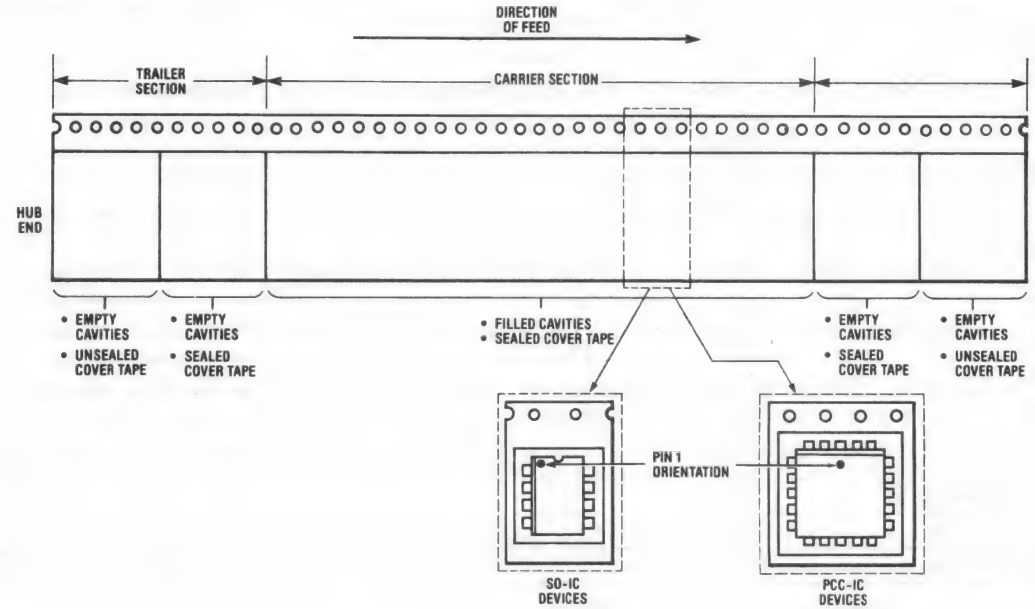
→ Direction of Feed

	Trailer (Hub End)*		Carrier*	Leader (Start End)*	
	Empty Cavities, min (Unsealed Cover Tape)	Empty Cavities, min (Sealed Cover Tape)	Filled Cavities (Sealed Cover Tape)	Empty Cavities, min (Sealed Cover Tape)	Empty Cavities, min (Unsealed Cover Tape)
Small Outline IC					
SO-8 (Narrow)	2	2	2500	5	5
SO-14 (Narrow)	2	2	2500	5	5
SO-14 (Wide)	2	2	1000	5	5
SO-16 (Narrow)	2	2	2500	5	5
SO-16 (Wide)	2	2	1000	5	5
SO-20 (Wide)	2	2	1000	5	5
SO-24 (Wide)	2	2	1000	5	5
Plastic Chip Carrier IC					
PCC-20	2	2	1000	5	5
PCC-28	2	2	750	5	5
PCC-44	2	2	500	5	5

*The following diagram identifies these sections of the tape and Pin #1 device orientation.

Short-Form Procurement Specification (Continued)

DEVICE ORIENTATION

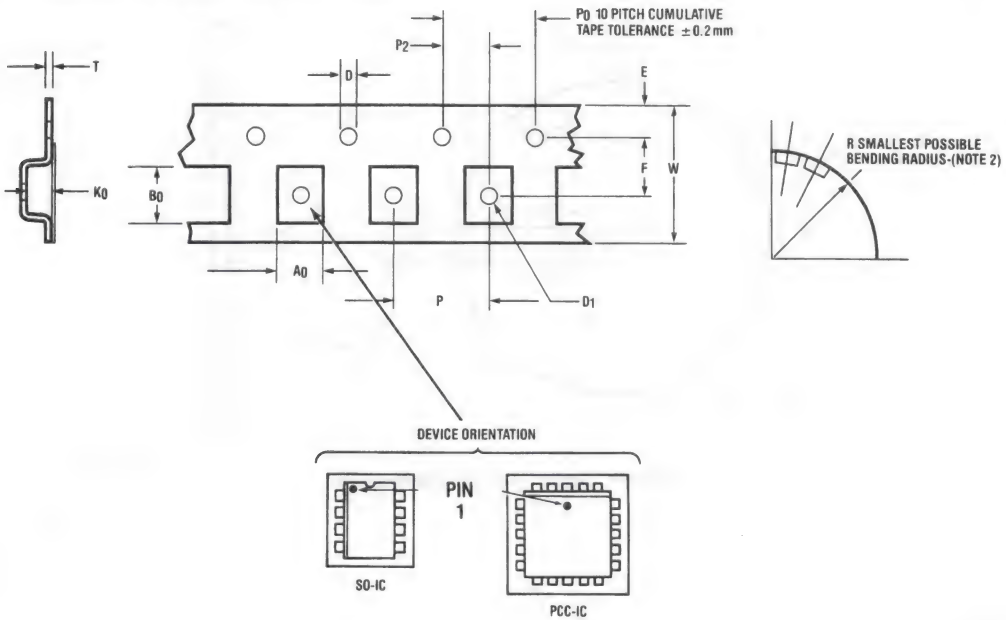


TL/XX/0026–8

MATERIALS

- Cavity Tape: Conductive PVC (less than 10^5 Ohms/Sq)
- Cover Tape: Polyester
  - (1) Conductive cover available
- Reel:
  - (1) Solid 80 pt fibreboard (standard)
  - (2) Conductive fibreboard available
  - (3) Conductive plastic (PVC) available

TAPE DIMENSIONS (24 Millimeter Tape or Less)



TL/XX/0026–9

Short-Form Procurement Specification (Continued)

	W	P	F	E	P_2	P_0	D	T	A_0	B_0	K_0	D_1	R
Small Outline IC													
SO-8 (Narrow)	12±.30	8.0±.10	5.5±.05	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	6.4±.10	5.2±.10	2.1±.10	1.55±.05	30
SO-14 (Narrow)	16±.30	8.0±.10	7.5±.10	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	6.5±.10	9.0±.10	2.1±.10	1.55±.05	40
SO-14 (Wide)	16±.30	12.0±.10	7.5±.10	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	10.9±.10	9.5±.10	3.0±.10	1.55±.05	40
SO-16 (Narrow)	16±.30	8.0±.10	7.5±.10	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	6.5±.10	10.3±.10	2.1±.10	1.55±.05	40
SO-16 (Wide)	16±.30	12.0±.10	7.5±.10	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	10.9±.10	10.76±.10	3.0±.10	1.55±.05	40
SO-20 (Wide)	24±.30	12.0±.10	11.5±.10	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	10.9±.10	13.3±.10	3.0±.10	2.05±.05	50
SO-24 (Wide)	24±.30	12.0±.10	11.5±.10	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	10.9±.10	15.85±.10	3.0±.10	2.05±.05	50
Plastic Chip Carrier IC													
PCC-20	16±.30	12.0±.10	7.5±.10	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	9.3±.10	9.3±.10	4.9±.10	1.55±.05	40
PCC-28	24±.30	16.0±.10	11.5±.10	1.75±.10	2.0±.05	4.0±.10	1.55±.05	.30±.10	13.0±.10	13.0±.10	4.9±.10	2.05±.05	50

Note 1: A_0, B_0 and K_0 dimensions are measured 0.3 mm above the inside wall of the cavity bottom.

Note 2: Tape with components shall pass around a mandril radius R without damage.

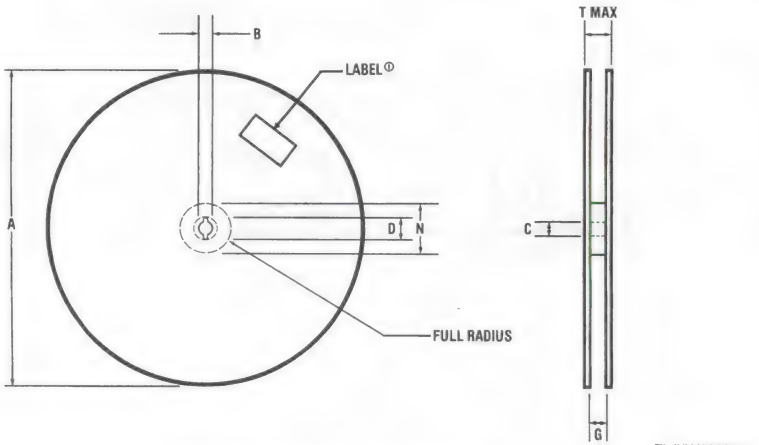
Note 3: Cavity tape material shall be PVC conductive (less than 10^5 Ohms/Sq).

Note 4: Cover tape material shall be polyester (30–65 grams peel-back force).

Note 5: D_1 Dimension is centered within cavity.

Note 6: All dimensions are in millimeters.

REEL DIMENSIONS



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STAR™* Surface Mount Tape and Reel

Short-Form Procurement Specifications (Continued)

		A (Max)	B (Min)	C	D (Min)	N (Min)	G	T (Max)
12 mm Tape	SO-8 (Narrow)	(13.00) (330)	.059 1.5	.512±.002 13±0.05	.795 20.2	1.969 50	$0.488^{+.078}_{-.000}$ 12.4^{+2}_{-0}	.724 18.4
16 mm Tape	SO-14 (Narrow) SO-14 (Wide) SO-16 (Narrow) SO-16 (Wide) PCC-20	(13.00) (330)	.059 1.5	.512±.002 13±0.05	.795 20.2	1.969 50	$0.646^{+.078}_{-.000}$ 16.4^{+2}_{-0}	.882 22.4
24 mm Tape	SO-20 (Wide) SO-24 (Wide) PCC-28	(13.00) (330)	.059 1.5	.512±.002 13±0.05	.795 20.2	1.969 50	$0.960^{+.078}_{-.000}$ 24.4^{+2}_{-0}	1.197 30.4
32 mm Tape	PCC-44	(13.00) (330)	.059 1.5	.512±.002 13±0.05	.795 20.2	1.969 50	$1.276^{+.078}_{-.000}$ 32.4^{+2}_{-0}	1.512 38.4

Units: Inches / Millimeters

Material: Paperboard (Non-Flaking)

LABEL

Human and Machine Readable Label is provided on reel. A variable (C.P.I) density code 39 is available. NSC STD label (7.6 C.P.I.)

FIELD

Lot Number

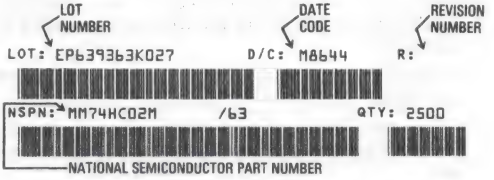
Date Code

Revision Level

National Part No. I.D.

Qty.

EXAMPLE



TL/XX/0026–11

Fields are separated by at least one blank space.

Future Tape-and-Reel packs will also include a smaller-size bar code label (high-density code 39) at the beginning of the tape. (This tape label is not available on current production.)

National Semiconductor will also offer additional labels containing information per your specific specification.

Wave Soldering of Surface Mount Components

ABSTRACT

In facing the upcoming surge of "surface mount technology", many manufacturers of printed circuit boards have taken steps to convert some portions of their boards to this new process. However, as the availability of surface mount components is still limited, may have taken to mixing the lead-inserted standard dual-in-line packages (DIPs) with the surface mounted devices (SMDs). Furthermore, to take advantage of using both sides of the board, surface-mounted components are generally adhered to the bottom side of the board while the top side is reserved for the conventional lead-inserted packages. If processed through a wave solder machine, the semiconductor components are now subjected to extra thermal stresses (now that the components are totally immersed into the molten solder).

A discussion of the effect of wave soldering on the reliability of plastic semiconductor packages follows. This is intended to highlight the limitations which should be understood in the use of wave soldering of surface mounted components.

ROLE OF WAVE-SOLDERING IN APPLICATION OF SMDs

The generally acceptable methods of soldering SMDs are vapor phase reflow soldering and IR reflow soldering, both requiring application of solder paste on PW boards prior to placement of the components. However, sentiment still exists for retaining the use of the old wave-soldering machine.

Wave Soldering of Surface Mount Components (Continued)

The reasons being:

1) Most PC Board Assembly houses already possess wave soldering equipment. Switching to another technology such as vapor phase soldering requires substantial investment in equipment and people.
2) Due to the limited number of devices that are surface mount components, it is necessary to mix both lead inserted components and surface mount components on the same board.
3) Some components such as relays and switches are made of materials which would not be able to survive the temperature exposure in a vapor phase or IR furnace.

PW BOARD ASSEMBLY PROCEDURES

There are two considerations in which through-hole ICs may be combined with surface mount components on the PW Board:

a) Whether to mount ICs on one or both sides of the board.
b) The sequence of soldering using Vapor Phase, IR or Wave Soldering singly or combination of two or more methods.

The various processes that may be employed are:

A) Wave Solder before Vapor/IR reflow solder.

1. Components on the same side of PW Board.

   Lead insert standard DIPS onto PW Board Wave solder (conventional)

   Wash and lead trim

   Dispense solder paste on SMD pads

   Pick and place SMDs onto PW Board

   Bake

   Vapor phase/IR reflow

   Clean

2. Components on opposite side of PW Board.

   Lead insert standard DIPs onto PW Board

   Wave Solder (conventional)

   Clean and lead trim

   Invert PW Board

   Dispense solder paste on SMD pads

   Dispense drop of adhesive on SMD sites (optional for smaller components)

   Pick and place SMDs onto board

   Bake/Cure

   Invert board to rest on raised fixture

   Vapor/IR reflow soldering

   Clean

B) Vapor/IR reflow solder then Wave Solder.

1. Components on the same side of PW Board.

   Solder paste screened on SMD side of Printed Wire Board

   Pick and place SMDs

   Bake

   Vapor/IR reflow

   Lead insert on same side as SMDs

   Wave solder

   Clean and trim underside of PCB

C) Vapor/IR reflow only.

1. Components on the same side of PW Board.

   Trim and form standard DIPs in "gull wing" configuration

   Solder paste screened on PW Board

   Pick and place SMDs and DIPs

   Bake

   Vapor/IR reflow

   Clean

2. Components on opposite sides of PW Board.

   Solder paste screened on SMD-side of Printed Wire Board

   Adhesive dispensed at central location of each component

   Pick and place SMDs

   Bake

   Solder paste screened on all pads on DIP-side or alternatively apply solder rings (performs) on leads

   Lead insert DIPs

   Vapor/IR reflow

   Clean and lead trim

D) Wave Soldering Only

1. Components on opposite sides of PW Board.

   Adhesive dispense on SMD side of PW Board

   Pick and place SMDs

   Cure adhesive

   Lead insert top side with DIPs

   Wave solder with SMDs down and into solder bath

   Clean and lead trim

All of the above assembly procedures can be divided into three categories for I.C. Reliability considerations:

1) Components are subjected to both a vapor phase/IR heat cycle then followed by a wave-solder heat cycle or vice versa.
2) Components are subjected to only a vapor phase/IR heat cycle.
3) Components are subjected to wave-soldering only and SMDs are subjected to heat by immersion into a solder pot.

Of these three categories, the last is the most severe regarding heat treatment to a semiconductor device. However, note that semiconductor molded packages generally possess a coating of solder on their leads as a final finish for solderability and protection of base leadframe material. Most semiconductor manufacturers solder-plate the component leads, while others perform hot solder dip. In the latter case the packages may be subjected to total immersion into a hot solder bath under controlled conditions (manual operation) or be partially immersed while in a 'pallet' where automatic wave or DIP soldering processes are used. It is, therefore, possible to subject SMDs to solder heat under certain conditions and not cause catastrophic failures.

Wave Soldering of Surface Mount Components (Continued)

THERMAL CHARACTERISTICS OF MOLDED INTEGRATED CIRCUITS

Since Plastic DIPs and SMDs are encapsulated with a thermoset epoxy, the thermal characteristics of the material generally correspond to a TMA (Thermo-Mechanical Analysis) graph. The critical parameters are (a) its Linear thermal expansion characteristics and (b) its glass transition temperature after the epoxy has been fully cured. A typical TMA graph is illustrated in *Figure 1*. Note that the epoxy changes to a higher thermal expansion once it is subjected to temperatures exceeding its glass transition temperature. Metals (as used on lead frames, for example) do not have this characteristic and generally will have a consistent Linear thermal expansion over the same temperature range.

In any good reliable plastic package, the choice of lead frame material should be such to match its thermal expansion properties to that of the encapsulating epoxy. In the event that there is a mismatch between the two, stresses can build up at the interface of the epoxy and metal. There now exists a tendency for the epoxy to separate from the metal lead frame in a manner similar to that observed on bimetallic thermal range.

In most cases when the packages are kept at temperatures below their glass transition, there is a small possibility of separation at the expoxy-metal interface. Howerver, if the package is subjected to temprature above its glass-transition temperature, the epoxy will begin to expand much faster than the metal and the probability of separation is greatly increased.

CONVENTIONAL WAVE-SOLDERING

Most wave-soldering operations occur at temperatures between 240–260°C. Conventional epoxies for encapsulation have glass-transition temperature between 140–170°C. An I.C. directly exposed to these temperatures risks its long term functionality due to epoxy/metal separation.

Fortunately, there are factors that can reduce that element of risk:

1) The PW board has a certain amount of heat-sink effect and tends to shield the components from the temperature of the solder (if they were placed on the top side of the board). In actual measurements, DIPs achieve a temperature between 120–150°C in a 5-second pass over the solder. This accounts for the fact that DIPs mounted in the conventional manner are reliable.
2) In conventional soldering, only the tip of each lead in a DIP would experience the solder temperature because the epoxy and die are standing above the PW board and out of the solder bath.

EFFECT ON PACKAGE PERFORMANCE BY EPOXY-METAL SEPARATION

In wave soldering, it is necessary to use fluxes to assist the solderability of the components and PW boards. Some facilities may even process the boards and components through some form of acid cleaning prior to the soldering temperature. If separation occurs, the flux residues and acid residues (which may be present owing to inadequate cleaning) will be forced into the package mainly by capillary action as the residues move away from the solder heat source. Once the package is cooled, these contaminants are now trapped within the package and are available to diffuse with moisture from the epoxy over time. It should be noted that electrical tests performed immediately after soldering generally will give no indication of this potential problem. In any case, the end result will be corrosion of the chip metallization over time and premature failure of the device in the field.

VAPOR PHASE/IR REFLOW SOLDERING

In both vapor phase and IR reflow soldering, the risk of separation between epoxy/metal can also be high. Operating temperatures are 215°C (vapor phase) or 240°C (IR) and duration may also be longer (30 sec–60 sec). On the same theoretical basis, there should also be separation. However, in both these methods, solder paste is applied to the pads of the boards; no fluxes are used. Also, the devices are not immersed into the hot solder. This reduces the possibility of solder forcing itself into the epoxy-lead frame interface. Furthermore, in the vapor phase system, the soldering environment is "oxygen-free" and considered "contaminant free". Being so, it could be visualized that as far as reliability with respect to corrosion, both of these methods are advantageous over wave soldering.

BIAS MOISTURE TEST

A bias moisture test was designed to determine the effect on package performance. In this test, the packages are pressured in a stream chamber to accelerate penetration of moisture into the package. An electrical bias is applied on the device. Should there be any contaminants trapped within the package, the moisture will quickly form an electrolyte and cause the electrodes (which are the lead fingers), the gold wire and the aluminum bond-pads of the silicon device to corrode. The aluminum bond-pads, being the weakest link of the system, will generally be the first to fail.

This proprietary accelerated bias/moisture pressure-test is significant in relation to the life test condition at 85°C and

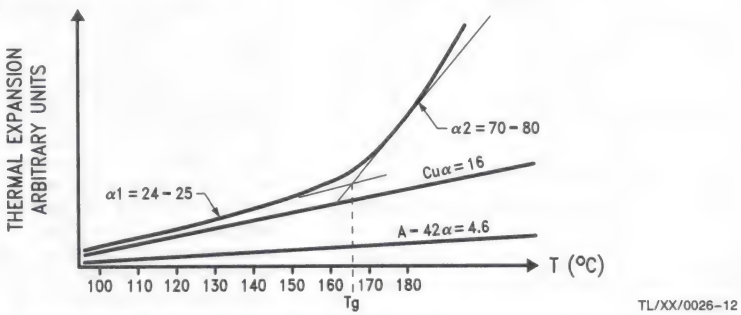


FIGURE 1. Thermal Expansion and Glass Transition Temperature

Wave Soldering of Surface Mount Components (Continued)

85% relative humidity. Once cycle of approximately 100 hours has been shown to be equivalent to 2000 hours in the 85/85 condition. Should the packages start to fail within the first cycle in the test, it is anticipated that the boards with these components in the harsh operating environment (85°C/85% RH) will experience corrosion and eventual electrical failures within its first 2000 hours of operation.

Whether this is significant to a circuit board manufacturer will obviously be dependent on the products being manufactured and the workmanship or reliability standards. Generally in systems with a long warranty and containing many components, it is advisable both on a reputation and cost basis to have the most reliable parts available.

TEST RESULTS

The comparison of vapor phase and wave-soldering upon the reliability of molded Small-Outline packages was performed using the bias moisture test (see Table IV). It is clearly seen that vapor phase reflow soldering gave more consistent results. Wave-soldering results were based on manual operation giving variations in soldering parameters such as temperature and duration.

TABLE IV. Vapor Phase vs. Wave Solder

1. Vapor phase (60 sec. exposure @ 215°C)
   = 9 failures/1723 samples
   = 0.5% (average over 32 sample lots)
2. Wave solder (2 sec total immersion @ 260°C)
   = 16 failures/1201 samples
   = 1.3% (average over 27 sample lots)

Package: SO-14 lead
Test: Bias moisture test 85% R.H., 85°C for 2000 hours
Device: LM324M

In Table V we examine the tolerance of the Small-Outlined (SOIC) package to varying immersion time in a hot solder pot. SO-14 lead molded packages were subjected to the bias moisture test after being treated to the various soldering conditions and repeated four (4) times. End point was an electrical test after an equivalent of 4000 hours 85/85 test. Results were compared for packages by itself against packages which were surface-mounted onto a FR-4 printed wire board.

TABLE V. Summary of Wave Solder Results (85% R.H./85°C Bias Moisture Test, 2000 hours) (# Failures/Total Tested)

	Unmounted	Mounted
Control/Vapor Phase 15 sec @ 215°C	0/114	0/84
Solder Dip 2 sec @ 260°C	2/144 (1.4%)	0/85
Solder Dip 4 sec @ 260°C	—	0/83
Solder Dip 6 sec @ 260°C	13/248 (5.2%)	1/76 (1.3%)
Solder Dip 10 sec @ 260°C	14/127 (11.0%)	3/79 (3.8%)
Package: SO-14 lead Device: LM324M		

Since the package is of very small mass and experiences a rather sharp thermal shock followed by stresses created by the mismatch in expansion, the results show the package being susceptible to failures after being immersed in excess of 6 seconds in a solder pot. In the second case where the packages were mounted, the effect of severe temperature excursion was reduced. In the second case where the packages were mounted, the effect of severe temperature excursion was reduced. In any case, because of the repeated treatment, the package had failures when subjected in excess of 6 seconds immersion in hot solder. The safety margin is therefore recommended as maximum 4 seconds immersion. If packages were immersed longer than 4 seconds, there is a probable chance of finding some long term reliability failures even though the immediate electrical test data could be acceptable.

Finally, Table VI examines the bias moisture test performed on surface mount (SOIC) components manufactured by various semiconductor houses. End point was an electrical test after an equivalent of 6000 hours in a 85/85 test. Failures were analyzed and corrosion was checked for in each case to detect flaws in package integrity.

TABLE VI. U.S. Manufacturers Integrated Circuits Reliability in Various Solder Environments (# Failure/Total Tested)

Package SO-8	Vapor Phase 30 sec	Wave Solder 2 sec	Wave Solder 4 sec	Wave Solder 6 sec	Wave Solder 10 sec
Manuf A	8/30*	1/30*	0.30	12/30*	16/30*
Manuf B	2/30*	8/30*	2/30*	22/30*	20/30*
Manuf C	0/30	0/29	0/29	0/30	0/30
Manuf D	1/30*	0/30	12/30*	14/30*	2/30*
Manuf E	1/30**	0/30	0/30	0/30	0/30
Manuf F	0/30	0/30	0/30	0/30	0/30
Manuf G	0/30	0/30	0/30	0/30	0/30

*Corrosion-failures

**No Visual Defects—Non-corrosion failures

Test: Accelerated Bias Moisture Test; 85% R.H./85°C, 6000 equivalent hours.

SUMMARY

Based on the results presented, it is noted that surface-mounted components are as reliable as standard molded DIP packages. Whereas DIPs were never processed by being totally immersed in a hot solder wave during printed circuit board soldering, surface mounted components such as SOICs (Small Outline) are expected to survive a total immersion in the hot solder in order to capitalize on maximum population on boards. Being constructed from a thermoset plastic of relatively low Tg compared to the soldering temperature, the ability of the package to survive is dependent on the time of immersion and also the cleanliness of material. The results indicate that one should limit the immersion time of package in the solder wave to a maximum of 4 seconds in order to truly duplicate the reliability of a DIP. As the package size is reduced, as in a SO-8 lead, the requirement becomes even more critical. This is shown by the various manufacturers' performance. Results indicate there is room for improvement since not all survived the hot solder immersion without compromise to lower reliability.

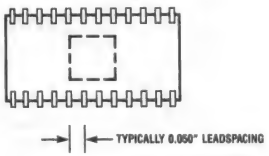
Small Outline (SO) Package Surface Mounting Methods-Parameters and Their Effect on Product Reliability

National Semiconductor
Application Note 450
Josip Huljev
W. K. Boey

The SO (small outline) package has been developed to meet customer demand for ever-increasing miniaturization and component density.

COMPONENT SIZE COMPARISON

S.O. Package



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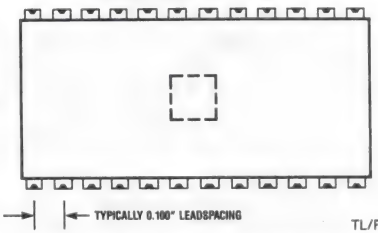
Standard DIP Package



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Because of its small size, reliability of the product assembled in SO packages needs to be carefully evaluated.

SO packages at National were internally qualified for production under the condition that they be of comparable reliability performance to a standard dual in line package under all accelerated environmental tests. *Figure A* is a summary of accelarated bias moisture test performance on 30V bipolar and 15V CMOS product assembled in SO and DIP (control) packages.

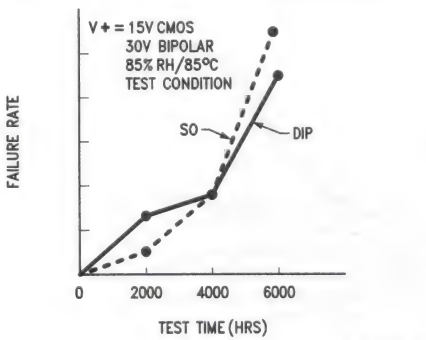

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FIGURE A

In order to achieve reliability performance comparable to DIPs—SO packages are designed and built with materials and processes that effectively compensate for their small size.

All SO packages tested on 85%RA, 85°C were assembled on PC conversion boards using vapor-phase reflow soldering. With this approach we are able to measure the effect of surface mounting methods on reliability of the process. As illustrated in *Figure A* no significant difference was detected between the long term reliability performance of surface mounted S.O. packages and the DIP control product for up to 6000 hours of accelerated 85%/85°C testing.

SURFACE-MOUNT PROCESS FLOW

The standard process flowcharts for basic surface-mount operation and mixed-lead insertion/surface-mount operations, are illustrated on the following pages.

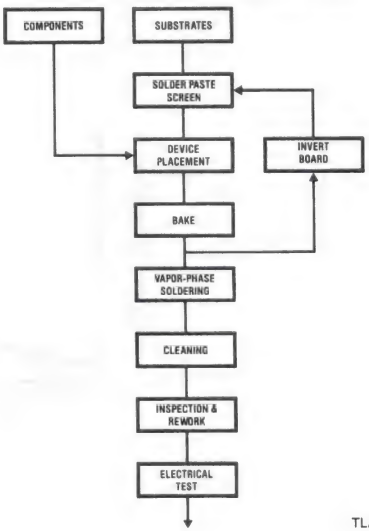
Usual variations encountered by users of SO packages are:

- Single-sided boards, surface-mounted components only.
- Single-sided boards, mixed-lead inserted and surface-mounted components.
- Double-sided boards, surface-mounted components only.
- Double-sided boards, mixed-lead inserted and surface-mounted components.

In consideration of these variations, it became necessary for users to utilize techniques involving wave soldering and adhesive applications, along with the commonly-used vapor-phase solder reflow soldering technique.

PRODUCTION FLOW

Basic Surface-Mount Production Flow



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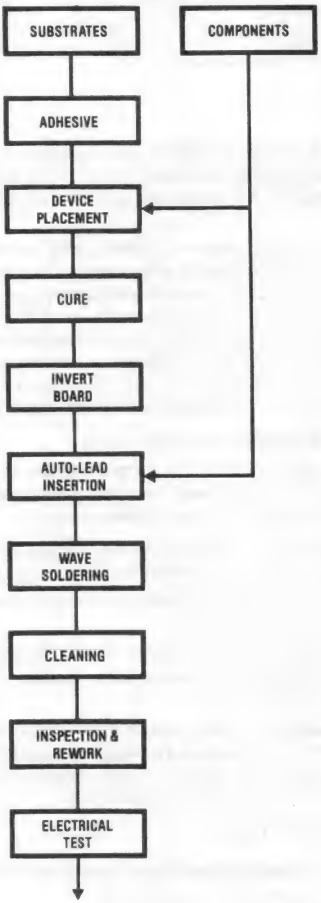
Mixed Surface-Mount and Axial-Leaded Insertion Components Production Flow



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Thermal stress of the packages during surface-mounting processing is more severe than during standard DIP PC board mounting processes. *Figure B* illustrates package temperature versus wave soldering dwell time for surface mounted packages (components are immersed into the molten solder) and the standard DIP wave soldering process. (Only leads of the package are immersed into the molten solder).

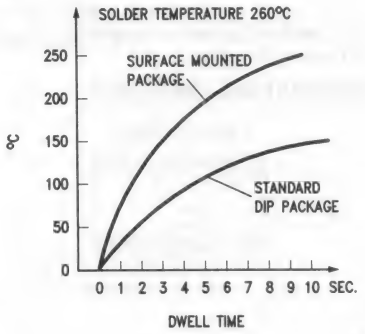

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FIGURE B

For an ideal package, the thermal expansion rate of the encapsulant should match that of the leadframe material in order for the package to maintain mechanical integrity during the soldering process. Unfortunately, a perfect matchup of thermal expansion rates with most presently used packaging materials is scarce. The problem lies primarily with the epoxy compound.

Normally, thermal expansion rates for epoxy encapsulant and metal lead frame materials are linear and remain fairly close at temperatures approaching 160°C, *Figure C.* At lower temperatures the difference in expansion rate of the two materials is not great enough to cause interface separation. However, when the package reaches the glass-transition temperature (T_g) of epoxy (typically 160–165°C), the thermal expansion rate of the encapsulant increases sharply, and the material undergoes a transition into a plastic state. The epoxy begins to expand at a rate three times or more greater than the metal leadframe, causing a separation at the interface.

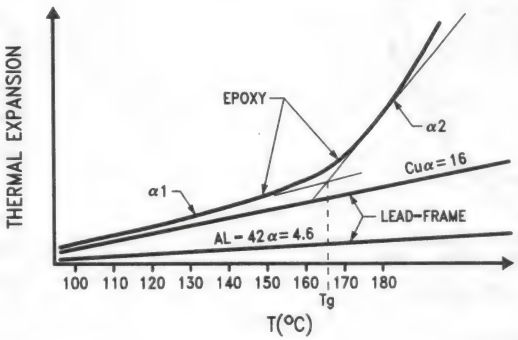

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FIGURE C

When this happens during a conventional wave soldering process using flux and acid cleaners, process residues and even solder can enter the cavity created by the separation and become entrapped when the material cools. These contaminants can eventually diffuse into the interior of the package, especially in the presence of moisture. The result is die contamination, excessive leakage, and even catastrophic failure. Unfortunately, electrical tests performed immediately following soldering may not detect potential flaws.

Most soldering processes involve temperatures ranging up to 260°C, which far exceeds the glass-transition temperature of epoxy. Clearly, circuit boards containing SMD packages require tighter process controls than those used for boards populated solely by DIPs.

Figure D is a summary of accelerated bias moisture test performance on the 30V bipolar process.

Group 1 — Standard DIP package

Group 2 — SO packages vapor-phase reflow soldered on PC boards

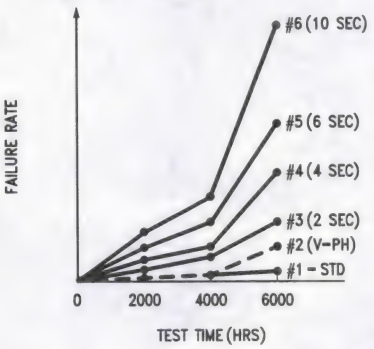
Group 3–6 SO packages wave soldered on PC boards

Group 3 — dwell time 2 seconds

4 — dwell time 4 seconds

5 — dwell time 6 seconds

6 — dwell time 10 seconds



TL/F/8766–7

FIGURE D

It is clear based on the data presented that SO packages soldered onto PC boards with the vapor phase reflow process have the best long term bias moisture performance and this is comparable to the performance of standard DIP packages. The key advantage of reflow soldering methods is the clean environment that minimized the potential for contamination of surface mounted packages, and is preferred for the surface-mount process.

When wave soldering is used to surface mount components on the board, the dwell time of the component under molten solder should be no more than 4 seconds, preferrably under 2 seconds in order to prevent damage to the component. Non-Halide, or (organic acid) fluxes are highly recommended.

PICK AND PLACE

The choice of automatic (all generally programmable) pick-and-place machines to handle surface mounting has grown considerably, and their selection is based on individual needs and degree of sophistication.

The basic component-placement systems available are classified as:

(a) In-line placement

— Fixed placement stations

— Boards indexed under head and respective components placed

(b) Sequential placement

— Either a X-Y moving table system or a θ, X-Y moving pickup system used

—Individual components picked and placed onto boards

(c) Simultaneous placement

— Multiple pickup heads

— Whole array of components placed onto the PCB at the same time

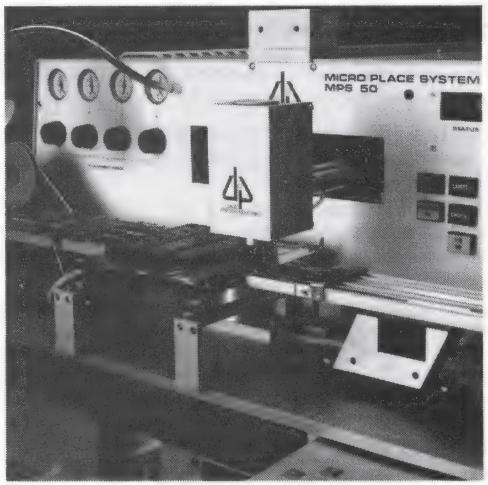
(d) Sequential/simultaneous placement

— X–Y moving table, multiple pickup heads system

— Components placed on PCB by successive or simultaneous actuation of pickup heads

The SO package is treated almost the same as surface-mount, passive components requiring correct orientation in placement on the board.

Pick and Place Action



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BAKE

This is recommended, despite claims made by some solder paste suppliers that this step be omitted.

The functions of this step are:

- Holds down the solder globules during subsequent reflow soldering process and prevents expulsion of small solder balls.
- Acts as an adhesive to hold the components in place during handling between placement to reflow soldering.
- Holds components in position when a double-sided surface-mounted board is held upside down going into a vapor-phase reflow soldering operation.
- Removes solvents which might otherwise contaminate other equipment.
- Initiates activator cleaning of surfaces to be soldered.
- Prevents moisture absorption.

The process is moreover very simple. The usual schedule is about 20 minutes in a 65°C–95°C (dependent on solvent system of solder paste) oven with adequate venting. Longer bake time is not recommended due to the following reasons:

- The flux will degrade and affect the characteristics of the paste.
- Solder globules will begin to oxidize and cause solderability problems.
- The paste will creep and after reflow, may leave behind residues between traces which are difficult to remove and vulnerable to electro-migration problems.

REFLOW SOLDERING

There are various methods for reflowing the solder paste, namely:

- Hot air reflow
- Infrared heating (furnaces)
- Convectional oven heating
- Vapor-phase reflow soldering
- Laser soldering

For SO applications, hot air reflow/infrared furnace may be used for low-volume production or prototype work, but vapor-phase soldering reflow is more efficient for consistency and speed. Oven heating is not recommended because of "hot spots" in the oven and uneven melting may result. Laser soldering is more for specialized applications and requires a great amount of investment.

HOT GAS REFLOW/INFRARED HEATING

A hand-held or table-mount air blower (with appropriate orifice mask) can be used.

The boards are preheated to about 100°C and then subjected to an air jet at about 260°C. This is a slow process and results may be inconsistent due to various heat-sink properties of passive components.

Use of an infrared furnace is the next step to automating the concept, except that the heating is promoted by use of IR lamps or panels. The main objection to this method is that certain materials may heat up at different rates under IR radiation and may result in damage to these components (usually sockets and connectors). This could be minimized by using far-infrared (non-focused) system.

VAPOR-PHASE REFLOW SOLDERING

Currently the most popular and consistent method, vapor-phase soldering utilizes a fluoroinert fluid with excellent heat-transfer properties to heat up components until the solder paste reflows. The maximum temperature is limited by the vapor temperature of the fluid.

The commonly used fluids (supplied by 3M Corp) are:

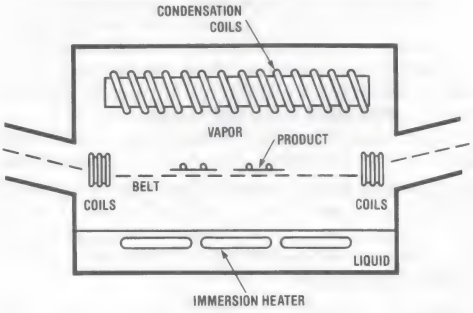
- FC-70, 215°C vapor (most applications) or FX-38
- FC-71, 253°C vapor (low-lead or tin-plate)

HTC, Concord, CA, manufactures equipment that utilizes this technique, with two options:

- Batch systems, where boards are lowered in a basket and subjected to the vapor from a tank of boiling fluid.
- In-line conveyorized systems, where boards are placed onto a continuous belt which transports them into a concealed tank where they are subjected to an environment of hot vapor.

Dwell time in the vapor is generally on the order of 15–30 seconds (depending on the mass of the boards and the loading density of boards on the belt).

In-Line Conveyorized Vapor-Phase Soldering



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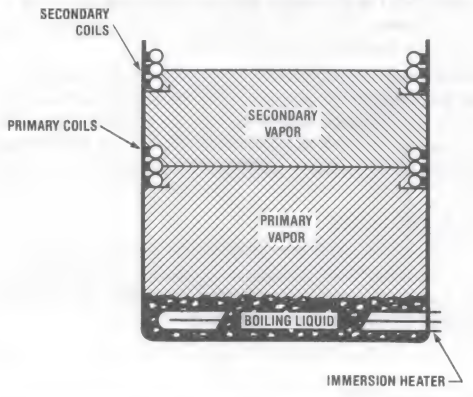
The question of thermal shock is asked frequently because of the relatively sharp increase in component temperature from room temperature to 215°C. SO packages mounted on representative boards have been tested and have shown little effect on the integrity of the packages. Various packages, such as cerdips, metal cans and TO-5 cans with glass seals, have also been tested.

Vapor-Phase Furnace

TL/F/8766–10

Batch-Fed Production Vapor-Phase Soldering Unit



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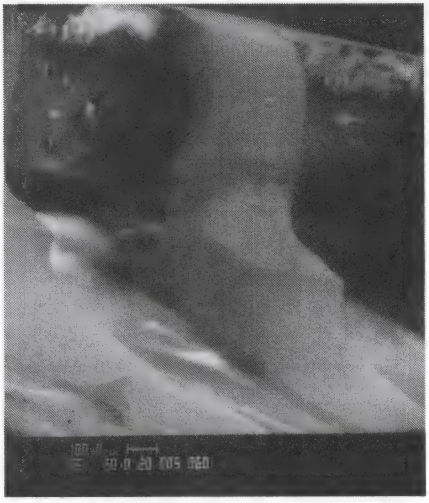
Solder Joints on a SO-14 Package on PCB

TL/F/8766–12

Solder Joints on a SO-14 Package on PCB



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PRINTED CIRCUIT BOARD

The SO package is molded out of clean, thermoset plastic compound and has no particular compatibility problems with most printed circuit board substrates.

The package can be reliably mounted onto substrates such as:

- G10 or FR4 glass/resin
- FR5 glass/resin systems for high-temperature applications
- Polymide boards, also high-temperature applications
- Ceramic substrates

General requirements for printed circuit boards are:

- Mounting pads should be solder-plated whenever applicable.
- Solder masks are commonly used to prevent solder bridging of fine lines during soldering.

  The mask also protects circuits from processing chemical contamination and corrosion.

  If coated over pre-tinned traces, residues may accumulate at the mask/trace interface during subsequent reflow, leading to possible reliability failures.

  Recommended application of solder resist on bare, clean traces prior to coating exposed areas with solder.

  General requirements for solder mask:

  — Good pattern resolution.

  — Complete coverage of circuit lines and resistance to flaking during soldering.

  — Adhesion should be excellent on substrate material to keep off moisture and chemicals.

  — Compatible with soldering and cleaning requirements.

SOLDER PASTE SCREEN PRINTING

With the initial choice of printed circuit lithographic design and substrate material, the first step in surface mounting is the application of solder paste.

The typical lithographic "footprints" for SO packages are illustrated below. Note that the 0.050″ lead center-center spacing is not easily managed by commercially-available air pressure, hand-held dispensers.

Using a stainless-steel, wire-mesh screen stencilled with an emulsion image of the substrate pads is by far the most common and well-tried method. The paste is forced through the screen by a V-shaped plastic squeegee in a sweeping manner onto the board placed beneath the screen.

The setup for SO packages has no special requirement from that required by other surface-mounted, passive components. Recommended working specifications are:

- Use stainless-steel, wire-mesh screens, #80 or #120, wire diameter 2.6 mils. Rule of thumb: mesh opening should be approximately 2.5–5 times larger than the average particle size of paste material.
- Use squeegee of Durometer 70.
- Experimentation with squeegee travel speed is recommended, if available on machine used.
- Use solder paste of mesh 200–325.
- Emulsion thickness of 0.005″ usually used to achieve a solder paste thickness (wet) of about 0.008″ typical.
- Mesh pattern should be 90 degrees, square grid.
- Snap-off height of screen should not exceed 1/8″, to avoid damage to screens and minimize distortion.

SOLDER PASTE

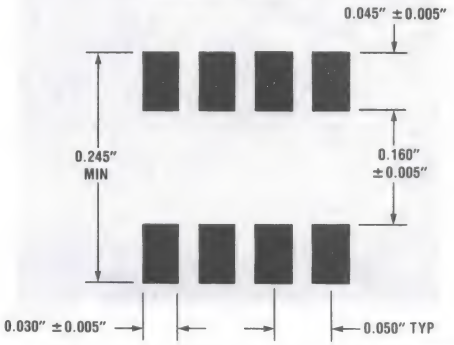
Selection of solder paste tends to be confusing, due to numerous formulations available from various manufacturers. In general, the following guidelines are sufficient to qualify a particular paste for production:

- Particle sizes (see photographs below). Mesh 325 (approximately 45 microns) should be used for general purposes, while larger (solder globules) particles are preferred for leadless components (LCC). The larger particles can easily be used for SO packages.

- Uniform particle distribution. Solder globules should be spherical in shape with uniform diameters and minimum amount of elongation (visual under 100/200 × magnification). Uneven distribution causes uneven melting and subsequent expulsion of smaller solder balls away from their proper sites.
- Composition, generally 60/40 or 63/37 Sn/Pb. Use 62/36 Sn/Pb with 2% Ag in the presence of Au on the soldering area. This formulation reduces problems of metal leaching from soldering pads.
- RMA flux system usually used.
- Use paste with aproximately 88–90% solids.

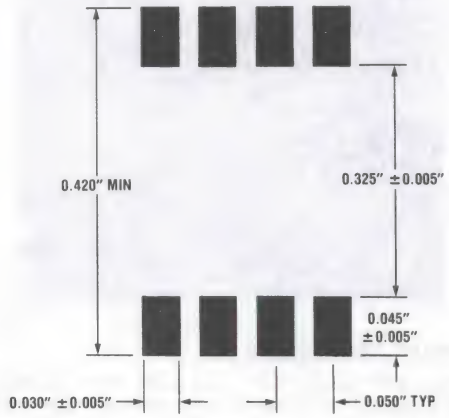
RECOMMENDED SOLDER PADS FOR SO PACKAGES

SO-8, SO-14, SO-16



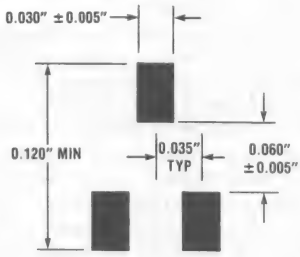
TL/F/8766–14

SO-16L, SO-20



TL/F/8766–15

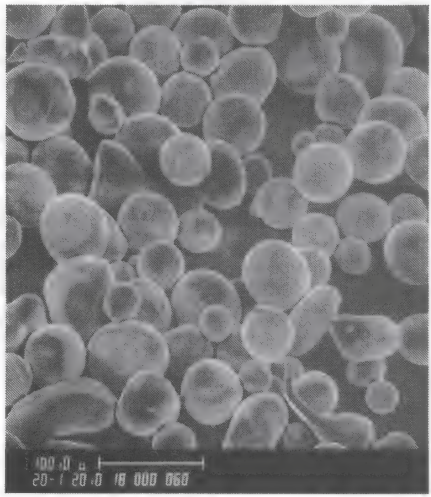
SOT-23



TL/F/8766–16

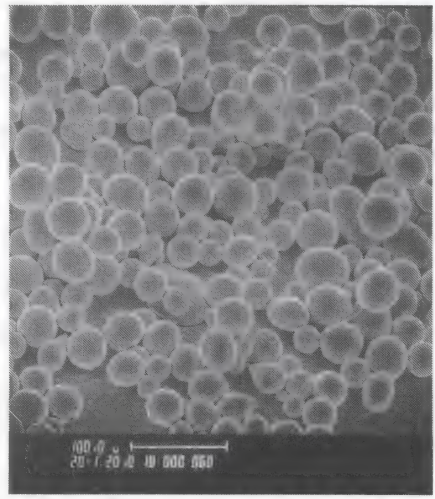
Comparison of Particle Size/Shape of Various Solder Pastes

200 × Alpha (62/36/2)



TL/F/8766–17

200 × Kester (63/37)



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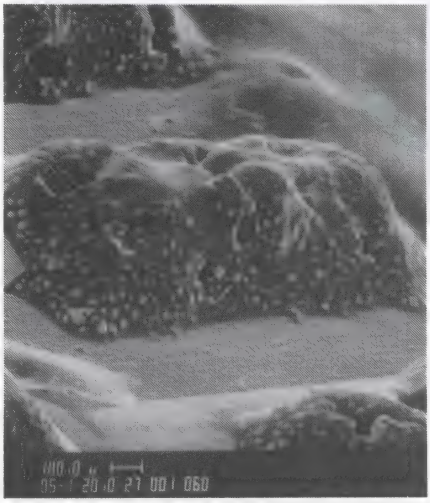
Comparison of Particle Size/Shape of Various Solder Pastes (Continued)

Solder Paste Screen on Pads



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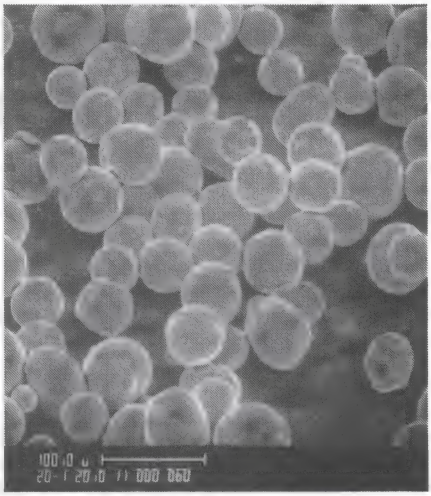
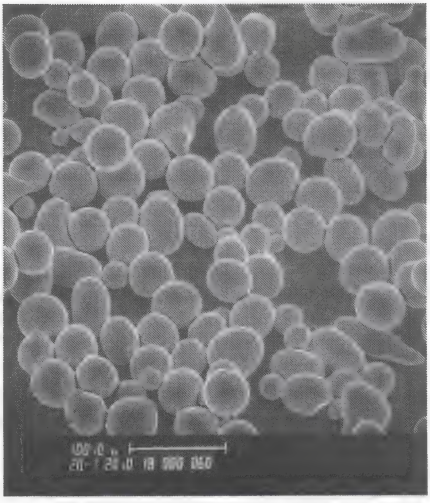
200 × Fry Metal (63/37)



TL/F/8766–20

200 ESL (63/37)



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CLEANING

The most critical process in surface mounting SO packages is in the cleaning cycle. The package is mounted very close to the surface of the substrate and has a tendency to collect residue left behind after reflow soldering.

Important considerations in cleaning are:

- Time between soldering and cleaning to be as short as possible. Residue should not be allowed to solidify on the substrate for long periods of time, making it difficult to dislodge.
- A low surface tension solvent (high penetration) should be employed. Solvents commercially available are:

  Freon TMS (general purpose)
  Freon TE35/TP35 (cold-dip cleaning)
  Freon TES (general purpose)

  It should also be noted that these solvents generally will leave the substrate surface hydrophobic (moisture repellent), which is desirable.

  Prelete or 1,1,1-Trichloroethane
  Kester 5120/5121
- A defluxer system which allows the workpiece to be subjected to a solvent vapor, followed by a rinse in pure solvent and a high-pressure spray lance are the basic requirments for low-volume production.
- For volume production, a conveyorized, multiple hot solvent spray/jet system is recommended.
- Rosin, being a natural occurring material, is not readily soluble in solvents, and has long been a stumbling block to the cleaning process. In recent developments, synthetic flux (SA flux), which is readily soluble in Freon TMS solvent, has been developed. This should be explored where permissible.

The dangers of an inadequate cleaning cycle are:

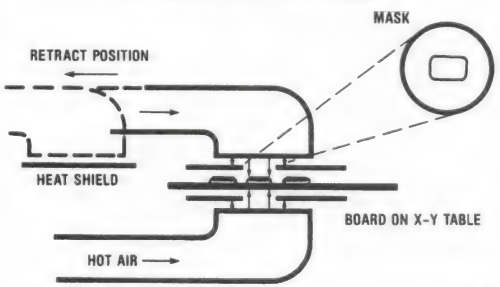
- Ion contamination, where ionic residue left on boards would cause corrosion to metallic components, affecting the performance of the board.
- Electro-migration, where ionic residue and moisture present on electrically-biased boards would cause dentritic growth between close spacing traces on the substrate, resulting in failures (shorts).

REWORK

Should there be a need to replace a component or re-align a previously disturbed component, a hot air system with appropriate orifice masking to protect surrounding components may be used.

When rework is necessary in the field, specially-designed tweezers that thermally heat the component may be used to remove it from its site. The replacement can be fluxed at the lead tips or, if necessary, solder paste can be dispensed onto the pads using a varimeter. After being placed into position, the solder is reflowed by a hot-air jet or even a standard soldering iron.

Hot-Air Solder Rework Station



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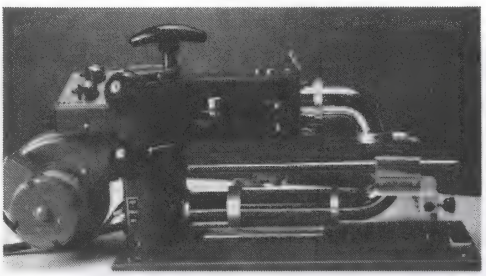
Hot-Air Rework Machine



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WAVE SOLDERING

In a case where lead insertions are made on the same board as surface-mounted components, there is a need to include a wave-soldering operation in the process flow.

Two options are used:

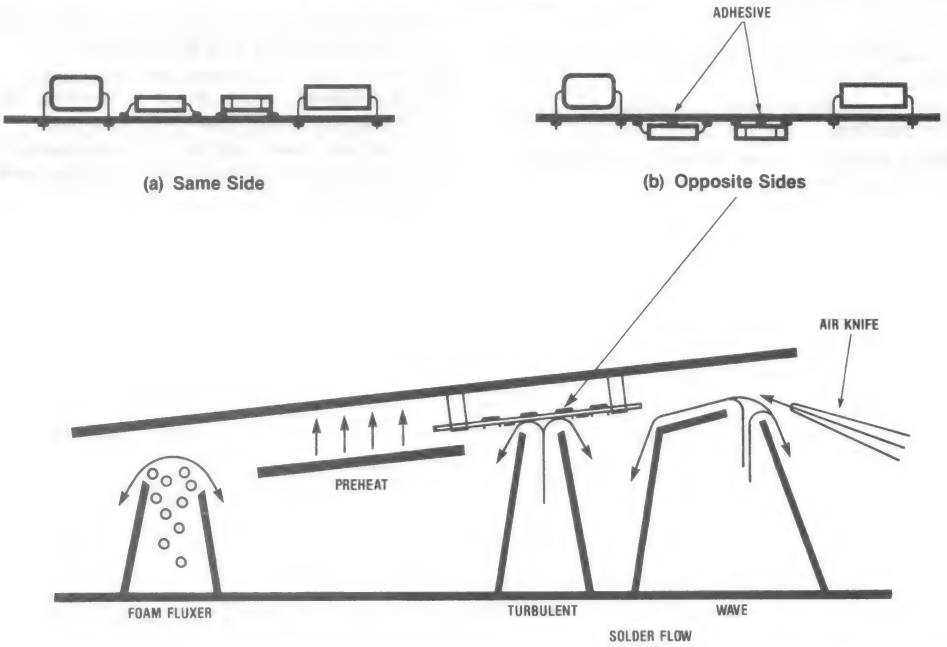
- Surface mounted components are placed and vapor phase reflowed before auto-insertion of remaining components. The board is carried over a standard wave-solder system and the underside of the board (only lead-inserted leads) soldered.
- Surface-mounted components are placed in position, but no solder paste is used. Instead, a drop of adhesive about 5 mils maximum in height with diameter not exceeding 25% width of the package is used to hold down the package. The adhesive is cured and then proceeded to auto-insertion on the reverse side of the board (surface-mounted side facing down). The assembly is then passed over a "dual wave" soldering system. Note that the surface-mounted components are immersed into the molten solder.

Lead trimming will pose a problem after soldering in the latter case, unless the leads of the insertion components are pre-trimmed or the board specially designed to localize certain areas for easy access to the trim blade.

The controls required for wave soldering are:

- Solder temperature to be 240–260°C. The dwell time of components under molten solder to be short (preferably kept under 2 seconds), to prevent damage to most components and semiconductor devices.
- RMA (Rosin Mildly Activated) flux or more aggressive OA (Organic Acid) flux are applied by either dipping or foam fluxing on boards prior to preheat and soldering. Cleaning procedures are also more difficult (aqueous, when OA flux is used), as the entire board has been treated by flux (unlike solder paste, which is more or less localized). Non-halide OA fluxes are highly recommended.
- Preheating of boards is essential to reduce thermal shock on components. Board should reach a temperature of about 100°C just before entering the solder wave.
- Due to the closer lead spacings (0.050″ vs 0.100″ for dual-in-line packages), bridging of traces by solder could occur. The reduced clearance between packages also causes "shadowing" of some areas, resulting in poor solder coverage. This is minimized by dual-wave solder systems.

Mixed Surface Mount and Lead Insertion



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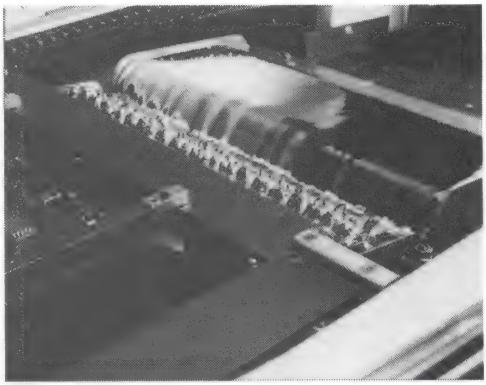
A typical dual-wave system is illustrated below, showing the various stages employed. The first wave typically is in turbulence and given a transverse motion (across the motion of the board). This covers areas where "shadowing" occurs. A second wave (usually a broad wave) then proceeds to perform the standard soldering. The departing edge from the solder is such to reduce "icicles," and is still further reduced by an air knife placed close to the final soldering step. This air knife will blow off excess solder (still in the fluid stage) which would otherwise cause shorts (bridging) and solder bumps.

AQUEOUS CLEANING

- For volume production, a conveyorized system is often used with a heated recirculating spray wash (water temperature 130°C), a final spray rinse (water temperature 45–55°C), and a hot (120°C) air/air-knife drying section.
- For low-volume production, the above cleaning can be done manually, using several water rinses/tanks. Fast-drying solvents, like alcohols that are miscible with water, are sometimes used to help the drying process.
- Neutralizing agents which will react with the corrosive materials in the flux and produce material readily soluble in water may be used; the choice depends on the type of flux used.
- Final rinse water should be free from chemicals which are introduced to maintain the biological purity of the water. These materials, mostly chlorides, are detrimental to the assemblies cleaned because they introduce a fresh amount of ionizable material.

Dual Wave



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CONFORMAL COATING

Conformal coating is recommended for high-reliability PCBs to provide insulation resistance, as well as protection against contamination and degradation by moisture.

Requirements:

- Complete coating over components and solder joints.
- Thixotropic material which will not flow under the packages or fill voids, otherwise will introduce stress on solder joints on expansion.
- Compatibility and possess excellent adhesion with PCB material/components.
- Silicones are recommended where permissible in application.

SMD Lab Support

FUNCTIONS

Demonstration—Introduce first-time users to surface-mounting processes.

Service—Investigate problems experienced by users on surface mounting.

Reliability Builds—Assemble surface-mounted units for reliability data acquisition.

Techniques—Develop techniques for handling different materials and processes in surface mounting.

Equipment—In conjunction with equipment manufacturers, develop customized equipments to handle high density, new technology packages developed by National.

In-House Expertise—Availability of in-house expertise on semiconductor research/development to assist users on packaging queries.

Section 9

Appendices/ Physical Dimensions

Section 9 Contents

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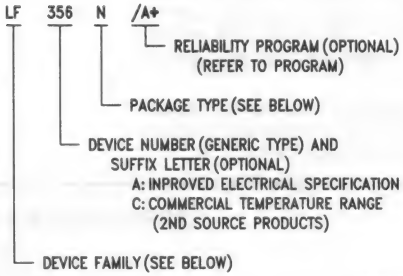
Distributors

National Semiconductor

Appendix A
General Product Marking & Code Explanation



TL/XX/0027–1

Device Family

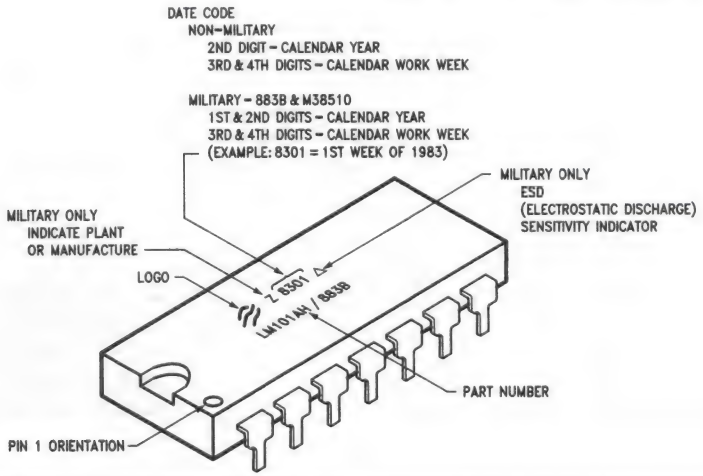
Integrated Circuits (IC's)

ADC	Data Conversion
AF	Active Filter
AH	Analog Switch (Hybrid)
AM	Analog Switch (Monolithic)
DAC	Data Conversion
DM	Digital (Monolithic)
HS	Hybrid
LF	Linear (Bifet)
LH	Linear (Hybrid)
LM	Linear (Monolithic)
LMC	Linear CMOS
LP	Linear (Low Power)
MF	Linear (Monolithic Filter)
SL	Special Linear
LMF	Linear Monolithic Filter

Package Type*

IC's Only

D	Glass/Metal DIP
E	Ceramic Leadless Chip Carrier (LCC)
F	Glass/Metal Flat Pak (¼" x ¼")
G	12 Lead TO-8 M/C
H	Multi-Lead M/C
H-05	4 Lead M/C (TO-5) } Shipped with
H-46	4 Lead M/C (TO-46) } Thermal Shield
J	Lo-Temp Ceramic DIP (Sometimes referred to as the "Fit-Seal" Package).
J-8	8 Lead Ceramic DIP ("MiniDIP")
J-14	14 Lead Ceramic DIP (-14 used only when product is also available in -8 pkg).
K	TO-3 M/C in Steel, except LM309K which is shipped in Aluminum
KC	TO-3 M/C (Aluminum)
K Steel	TO-3 M/C (Steel)
M	Small Outline Package
N	Molded DIP (EPOXY B)
N-01	Molded DIP (Epoxy B) with Staggered Leads
N-8	8 Lead Molded DIP (Epoxy B) ("Mini-DIP")
N-14	14 Lead Molded DIP (Epoxy B) (-14 used only when product is also available in -8 pkg).
P	3 Lead TO-202 PWR Pkg
Q	Cerdip with UV Window
T	3,5,11,15 & 23 Lead TO-220 PWR Pkg (Epoxy B)
V	Multi-lead Plastic Chip Carrier (PCC)
W	Lo-Temp Ceramic Flat Pak
WM	Wide Body Small Outline Package



TL/XX/0027–2

National Semiconductor **Appendix B**

APPLICATION NOTE REFERENCED BY PART NUMBER

National Semiconductor Linear Application notes are normally written to explain the operation and use of a particular device or family of IC's, or to present alternative technical solutions. The following PART NUMBER index references the published application notes that would offer application assistance for those specific IC's.

The 1986 Linear Applications Handbook is a complete text for all current Application Notes for both Monolithic and Hybrid products. Specific Application Notes are available upon request through National Semiconductor Sales Offices.

DEVICE NUMBER	APPLICATION NOTE
ADCXXXX	AN-156
ADC80	AN-360
ADC0801	AN-233, AN-271, AN-274, AN-280, AN-281, AN-294, LB-53
ADC0802	AN-233, AN-274, AN-280, AN-281, LB-53
ADC0803	AN-233, AN–274, AN-280, AN-281, LB-53
ADC0804	AN-233, AN-274, AN-276, AN-280, AN-281, LB-53
ADC0805	AN-233, AN-274, AN-280, AN-281, LB-53
ADC0808	AN-247, AN-280, AN-281
ADC0809	AN-247, AN-280
ADC0816	AN-193, AN-247, AN-258, AN-280
ADC0817	AN-247, AN-258, AN-280
ADC0820	AN-237
ADC0831	AN-280, AN-281
ADC0832	AN-280, AN-281
ADC0833	AN-280, AN-281
ADC0834	AN-280, AN-281
ADC0838	AN-280, AN-281
ADC1001	AN-276, AN-280, AN-281
ADC1005	AN-280
ADC1210	AN-245
ADC3501	AN-200, AN-202
ADC3511	AN-200
ADC3701	AN-200
ADC3711	AN-200
AH0014	AN-38
AH0019	AN-38
CD4016	AB-10
DACXXXX	AN-156
DAC0830	AN-284
DAC0831	AN-271, AN-284
DAC0832	AN-271, AN-284
DAC1000	AN-271, AN-275, AN-277, AN-284
DAC1001	AN-271, AN-275, AN-277, AN-284
DAC1002	AN-271, AN-275, AN-277, AN-284
DAC1006	AN-271, AN-275, AN-277, AN-284
DAC1007	AN-271, AN-275, AN-277, AN-284

DEVICE NUMBER	APPLICATION NOTE
DAC1008	AN-271, AN-275, AN-277, AN-284
DAC1020	AN-263, AN-269, AN-293, AN-294, AN-299
DAC1021	AN-269
DAC1022	AN-269
DAC1208	AN-271, AN-284
DAC1209	AN-271, AN-284
DAC1210	AN-271, AN-284
DAC1218	AN-293
DAC1220	AN-253, AN-269
DAC1221	AN-269
DAC1222	AN-269
DAC1230	AN-284
DAC1231	AN-271, AN-284
DAC1232	AN-271, AN-284
DAC1280	AN-261, AN-263
DH0034	AN-253
DH0035	AN-49
DS8606	AN-381, AN-382
DS8608	AN-382
DT1058	AN-287
DT1060	AN-287
DTSW250E2	AN-287
DTSW250GI	AN-287
INS8070	AN-260
LF111	LB-39
LF155	AN-263, AN-447
LF198	AN-245, AN-294
LF311	AN-301
LF347	AN-256, AN-262, AN-263, AN-265, AN-266, AN-301, AN-344, AN-447
LF351	AN-242, AN-263, AN-266, AN-271, AN-275, AN-293, AN-447, Appendix C
LF351A	AN-240
LF351B	Appendix D
LF353	AN-256, AN-258, AN-263, AN-264, AN-271, AN-285, AN-293, AN-447, LB-44, Appendix D
LF356	AN-253, AN-258, AN-260, AN-263, AN-266, AN-271, AN-272, AN-275, AN-293, AN-294, AN-295, AN-301, AN-447
LF357	AN-263, AN-447, LB-42
LF398	AN-247, AN-258, AN-266, AN-294, AN-298, LB-45
LF400	AN-428, AN-447
LF411	AN-294, AN-301, AN-344, AN-447
LF412	AN-272, AN-299, AN-301, AN-344, AN-447
LF441	AN-301, AN-447
LF13006	AN-344
LF13007	AN-344
LF13331	AN-294, AN-447
LF13508	AN-289, AN-360, AN-447
LF13509	AN-289, AN-295, AN-447
LH0002	AN-13, AN-63, AN-227, AN-244, AN-263, AN-272, AN-301
LH0022	AN-63, AN-75
LH0023	AN-245, AN-360
LH0024	AN-253
LH0032	AN-242, AN-244, AN-253
LH0033	AN-48, AN-115, AN-227, AN-253
LH0042	AN-63

DEVICE NUMBER	APPLICATION NOTE
LH0043	AN-245
LH0052	AN-63
LH0053	AN-245
LH0062	AN-75
LH0063	AN-227
LH0070	AN-301
LH0071	AN-245
LH0082	AN-244, AN-266
LH0086	AN-245, AN-360
LH0091	AN-180
LH0094	AN-301
LH0101	AN-261
LH1605	AN-343
LM10	AN-211, AN-247, AN-258, AN-271, AN-288, AN-299, AN-300
LM11	AN-241, AN-242, AN-260, AN-266, AN-271
LM12	AN-446
LM101	AN-4, AN-13, AN-20, AN-24, AN-75, LB-42, Appendix A
LM101A	AN-29, AN-30, AN-31, AN-79, AN-241, LB-1, LB-2, LB-4, LB-8, LB-14, LB-16, LB-19, LB-28
LM102	AN-4, AN-13, AN-30, LB-1, LB-5, LB-6, LB-11
LM103	AN-110, LB-41
LM104	AN-21, LB-3, LB-7, LB-10, LB-40
LM105	AN-21, AN-23, AN-110, LB-3, LB-7, LB-10
LM106	AN-41, LB-6, LB-12
LM107	AN-20, AN-31, LB-1, LB-12, LB-19, Appendix A
LM108	AN-29, AN-30, AN-31, AN-63, AN-79, AN-211, AN-241, LB-14, LB-15, LB-21
LM108A	AN-260, LB-15, LB-19
LM109	AN-42, LB-15
LM109A	LB-15
LM110	LB-11, LB-42
LM111	AN-41, AN-103, LB-12, LB-16, LB-32, LB-39
LM112	AN-63, LB-19
LM113	AN-56, AN-110, LB-21, LB-24, LB-28, LB-37
LM117	AN-178, AN-181, AN-182, LB-46, LB-47
LM117HV	LB-46, LB-47
LM118	LB-17, LB-19, LB-21, LB-23, Appendix A
LM119	AN-115, LB-23
LM120	AN-182
LM121	AN-79, AN-104, AN-184, AN-260, LB-22
LM121A	LB-32
LM122	AN-97, LB-38
LM125	AN-82
LM126	AN-82
LM129	AN-173,AN-178, AN-262, AN-266
LM131	AN-210, Appendix D
LM131A	AN-210
LM134	LB-41
LM135	AN-225, AN-262, AN-292, AN-298
LM137	LB-46
LM137HV	LB-46
LM138	LB-46
LM139	AN-74
LM143	AN-127, AN-271
LM148	AN-260

DEVICE NUMBER	APPLICATION NOTE
LM150	LB-46
LM158	AN-116
LM160	AN-87
LM161	AN-87, AN-266
LM163	AN-295
LM194	AN-222, LB-21
LM195	AN-110
LM199	AN-161, AN-260, AN-360
LM199A	AN-161
LM211	LB-39
LM216A	LB-37
LM231	AN-210
LM231A	AN-225
LM235	AN-225
LM239	AN-74
LM258	AN-116
LM260	AN-87
LM261	AN-87
LM301A	AN-178, AN-181, AN-222
LM304	LB-40
LM308	AN-88, AN-184, AN-272, LB-22, LB-28, Appendix D
LM308A	AN-225, LB-24
LM309	AN-178, AN-182
LM311	AN-41, AN-103, AN-260, AN-263, AN-288, AN-294, AN-295, AN-307, LB-12, LB-16, LB-18, LB-39
LM313	AN-263
LM316	AN-258
LM317	AN-178, LB-35, LB-46
LM317H	LB-47
LM318	AN-115, AN-299, LB-21
LM319	AN-115, AN-271, AN-293
LM320	AN-288
LM321	LB-24
LM324	AN-88, AN-258, AN-274, AN-284, AN-301, LB-44, AB-25, Appendix C
LM329	AN-256, AN-263, AN-284, AN-295, AN-301
LM329B	AN-225
LM330	AN-301
LM331	AN-210, AN-240, AN-265, AN-278, AN-285, AN-311, LB-45, Appendix, C Appendix D
LM331A	AN-210, Appendix C
LM334	AN-242, AN-256, AN-284
LM335	AN-225, AN-263, AN-295
LM336	AN-202, AN-247, AN-258
LM337	LB-46
LM338	LB-49, LB-51
LM339	AN-74, AN-245, AN-274
LM340	AN-103, AN-182
LM340L	AN-256
LM342	AN-288
LM346	AN-202, LB-54
LM347	LB-44
LM348	AN-202, LB-42
LM349	LB-42
LM358	AN-116, AN-247, AN-271, AN-274, AN-284, AN-298, Appendix C
LM358A	Appendix D

DEVICE NUMBER	APPLICATION NOTE
LM359	AN-278, AB-24
LM360	AN-87
LM361	AN-87, AN-294
LM363	AN-271
LM380	AN-69, AN-146
LM381	AN-64, AN-104
LM382	AN-147
LM385	AN-242, AN-256, AN-301, AN-344
LM386	LB-54
LM389	AN-256, AN-263, AN-264, AN-274
LM391	AN-272
LM392	AN-274, AN-286
LM393	AN-271, AN-274, AN-293
LM394	AN-262, AN-263, AN-264, AN-271, AN-293, AN-299, AN-311, LB-52
LM395	AN-178, AN-181, AN-262, AN-263, AN-266, AN-301, LB-28
LM399	AN-184
LM555	AB-7
LM556	AB-7
LM565	AN-46, AN-146
LM566	AN-146
LM567	AN-46
LM709	AN-24, AN-30
LM710	AN-41, LB-12
LM725	LB-22
LM741	AN-75, AN-79, LB-19, LB-22
LM832	AN-386, AN-390
LM833	AN-346
LM1036	AN-390
LM1310	AN-81
LM1524	AN-272, AN-288, AN-292, AN-293
LM1800	AN-81, AN-147
LM1812	AB-20
LM1818	AN-407
LM1820	LB-29
LM1823	AN-391
LM1828	Appendix B
LM1830	AB-10
LM1837	AN-407
LM1845	Appendix B
LM1863	AN-381, AN-382
LM1865	AN-382, AN-390
LM1870	AN-382
LM1886	AN-402
LM1889	AN-402
LM1894	AN-384, AN-386, AN-390
LM1897	AN-407
LM2878	AN-147
LM2889	AN-391, AN-402
LM2907	AN-162
LM2917	AN-162
LM2931	AB-12
LM2931CT	AB-11

DEVICE NUMBER	APPLICATION NOTE
LM3045	AN-286
LM3046	AN-146, AN-299
LM3089	AN-147
LM3524	AN-272, AN-288, AN-292, AN-293
LM3820	AN-147, LB-29
LM3900	AN-72, AN-263, AN-274, AN-278, LB-20, AB-24
LM3909	AN-154
LM3911	LB-27
LM3914	LB-48, AB-25
LM3915	AN-386
LM3999	AN161
LM4250	AN-88, LB-34
LM7800	AN-178
LM78L12	AN-146
LMC835	AN-435
LP324	AN-284
MF10	AN-307
MM1458	AN-116
MM1558	AN-116
MM1558C	AN-116
MM2716	LB-54
MM54104	AN-252, AN-287, LB-54
MM57110	AN-382
MM74C00	AN-88
MM74C02	AN-88
MM74C04	AN-88
MM74C948	AN-193
MM74LS138	LB-54
2N4339	AN-32
LH4101	AN-480
LM34/35	AN-460
LM32900	AN-478
LM3578	AB-30
LPXXXX	AN-462
LM34	AN-462
LM35	AN-462
LM385	AN-462
LMC13334	AN-462
LP2950	AN-462
LP2951	AN-462
LP311	AN-462
LP324	AN-462
LP339	AN-462
LP365	AN-462

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Appendix C
Summary of Commercial Reliability Programs

General

National Semiconductor Commercial Reliability Programs provide a broad range of off-the-shelf enhanced semiconductor products that supply an extra measure of quality and reliability needed in high-stress or difficult to service applications.

National's A+ and B+ programs allow each individual customer to:

- Minimize the need for incoming electrical inspection
- Eliminate the need and associated costs of using independent testing laboratories
- Reduction in infant mortality rate
- Reduction in reworked board costs
- Reduction in warranty and service costs

A+ Product Enhancement

The A+ Product Enhancement incorporates the benefits of the Multiple-Pass and Elevated Temperature along with "BURN-IN."

The A+ Program provides:

- 100% Temperature Cycling
- 100% Electrical Testing at Room and High Temperature
- 100% Burn-In Testing Combining Increased Temperature with Applied Voltage
- Acceptable Quality Levels Greater than Industry Norm

Typical A+ Flow is:

- SEM
- Assembly and Seal
- Four Hour 150°C Bake
- Five Temperature Cycles (0°C to +100°C)
- High Temperature Electrical Test
- Electrical Test
- Burn-In (160 hours at a minimum junction temperature of 125°C)
- DC Parametric and Function Tests
- Tightened Quality Control Inspection Plans

Note: Certain products may follow slightly different process flows dictated by specific capabilities and device characteristics, consult NSC.

P+ Product Enhancement

The P+ product enhancement program applies to regulator devices and offers an added advantage. P+ involves a dynamic self-heating burn-in that tests the thermal shutdown of the regulator. P+ is proven more effective than the standard 125°C burn-in as an early screen for infant mortality defects. It sharply reduces the cost of testing incoming components. Reliability Report L-140 further explains the P+ process. The following chart lists regulators which receive P+ prior to shipment and at no additional cost.

Device	Package Types				
	TO-3 K STEEL	TO-39 H	TO-220 T	TO-202 P	TO-92 Z
LM109/309	X	X			
LM117/317	X	X	X	X	
LM117HV/317HV	X	X			
LM120/320	X	X	X	X	
LM123/323	X				
LM137/337	X	X	X	X	
LM137HV/337HV	X	X			
LM138/338	X				
LM140/340	X	X	X	X	
LM145/345	X				
LM150/250/350	X				
LM196/396	X				
LM2930/2935/2940/2984			X		
LM2931			X		X
LM78XX			X		

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Appendix D
Military Aerospace Programs from National Semiconductor

This appendix is intended to provide a brief overview of military products available from National Semiconductor. For further information, refer to our *1987 Reliability Handbook.*

MIL-M-38510

The MIL-M-38510 Program, which is sometimes called the JAN IC Program, is administered by the Defense Electronics Supply Center (DESC). The purpose of this program is to provide the military community with standardized products that have been manufactured and screened to government-controlled specifications in government-certified facilities. All 38510 manufacturers must be formally qualified and their products listed on DESC's Qualified Products List (QPL) before devices can be marked and shipped as JAN product.

There are two processing levels specified within MIL-M-38510: Classes S and B. Class S is typically specified for space flight applications, while Class B is used for aircraft and ground systems. National is a major supplier of both classes of devices. Screening requirements are outlined in Table III.

Tables I and II explain the JAN device marking system.

Copies of MIL-M-38510, the QPL, and other related documents may be obtained from:

Naval Publications and Forms Center
5801 Tabor Avenue
Philadelphia, PA 19120
(212) 697-2179

DESC Specifications

DESC specifications are issued to provide standardized versions of devices which are not yet available as JAN product. MIL-STD-883 Class B screening is coupled with tightly controlled electrical specifications which have been written to allow a manufacturer to use his standard electrical tests. A current listing of National's DESC specification offerings can be obtained from our franchised distributors, sales offices, or DESC. DESC is located in Dayton, Ohio.

MIL-STD-883

Although originally intended to establish uniform test methods and procedures, MIL-STD-883 has also become the general specification for non-JAN military product. Revision C of this document defines the minimum requirements for a device to be marked and advertised as 883-compliant. Included are design and construction criteria, documentation controls, electrical and mechanical screening requirements, and quality control procedures. Details can be found in paragraph 1.2.1 of MIL-STD-883.

National offers both 883 Class B and 883 Class S product. The screening requirements for both classes of product are outlined in Table III.

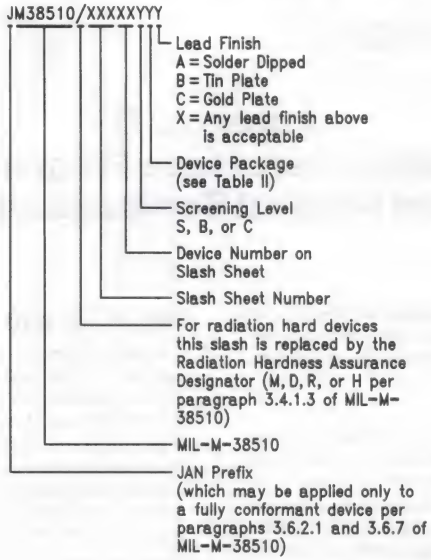
As with DESC specifications, a manufacturer is allowed to use his standard electrical tests provided that all critical parameters are tested. Also, the electrical test parameters, test conditions, test limits, and test temperatures must be clearly documented. At National Semiconductor, this information is available via our RETS (Reliability Electrical Test Specification Program). The RETS document is a complete description of the electrical tests performed and is controlled by our QA department. Individual copies are available upon request.

Some of National's older products are not completely compliant with MIL-STD-883 but are still required for use in military systems. These devices are screened to the same stringent requirements as 883 product but are marked "-MIL".

Military Screening Program (MSP)

National's Military Screening Program was developed to make screened versions of advanced products such as gate arrays and microprocessors available more quickly than is possible for JAN and 883 devices. Through this program, screened product is made available for prototypes and breadboards prior to or during the JAN or 883 qualification activities. MSP products receive the 100% screening of Table III but are not subjected to Group C and D quality conformance testing. Other criteria such as electrical testing and temperature range will vary depending upon individual device status and capability.

TABLE I. The MIL-M-38510 Part Marking


JM38510/XXXXXYYY
Lead Finish
A = Solder Dipped
B = Tin Plate
C = Gold Plate
X = Any lead finish above is acceptable
Device Package (see Table II)
Screening Level S, B, or C
Device Number on Slash Sheet
Slash Sheet Number
For radiation hard devices this slash is replaced by the Radiation Hardness Assurance Designator (M, D, R, or H per paragraph 3.4.1.3 of MIL-M-38510)
MIL-M-38510
JAN Prefix (which may be applied only to a fully conformant device per paragraphs 3.6.2.1 and 3.6.7 of MIL-M-38510)

CI24-1

TABLE II. JAN Package Codes

38510 Package Designation	Microcircuit Industry Description
A	14-Pin 1/4" X 1/4" (metal) flat pack
B	14-Pin 3/16" X 1/4" flat pack
C	14-Pin 1/4" X 3/4" dual-in-line
D	14-Pin 1/4" X 3/8" (ceramic) flat pack
E	16-Pin 1/4" X 3/8" dual-in-line
F	16-Pin 1/4" X 3/8" (metal or ceramic) flat pack
G	8-pin TO-99 can or header
H	10-pin 1/4" x 1/4" (metal) flat pack
I	10-pin TO-100 can or header
J	24-pin 1/2" x 1-1/4" dual-in-line
K	24-pin 3/8" x 5/8" flat pack
L	24-pin 1/4" x 1-1/4" dual-in-line
M	12-pin TO-101 can or header
N	(Note 1)
P	8-pin 1/4" x 3/8" dual-in-line
Q	40-pin 3/16" x 2-1/16" dual-in-line
R	20-pin 1/4" x 1-1/16" dual-in-line
S	20-pin 1/4" x 1/2" flat pack
T	(Note 1)
U	(Note 1)
V	18-pin 3/8" x 15/16" dual-in-line
W	22-pin 3/8" x 1-1/8" dual-in-line
X	(Note 1)
Y	(Note 1)
Z	(Note 1)
2	20-terminal 0.350" x 0.350" chip carrier
3	28-terminal 0.450" x 0.450" chip carrier

Note 1: These letters are assigned to packages by individual detail specifications and may be assigned to different packages in different specifications.

TABLE III. 100% Screening Requirements

Screen	Class S		Class B	
	Method	Reqmt	Method	Reqmt
1. Wafer Lot Acceptance	5007	All Lots		—
2. Nondestructive Bond Pull	2023	100%		—
3. Internal Visual (Note 1)	2010, Condition A	100%	2010, Condition B	100%
4. Stabilization Bake	1008, Condition C, 24 hrs. Min.	100%	1008, Condition C, 24 hrs. Min.	100%
5. Temp. Cycling (Note 2)	1010, Condition C	100%	1010, Condition C	100%
6. Constant Acceleration	2001, Condition E (Min.) Y_1 Orientation Only	100%	2001, Condition E, (Min.), Y_1 Orientation Only	100%
7. Visual Inspection (Note 3)		100%		100%
8. Particle Impact Noise Detection (PIND)	2020, Condition A (Note 4)	100%		—
9. Serialization	(Note 5)	100%		—
10. Interim (Pre-Burn-In) Electrical Parameters	Per Applicable Device Specification (Note 13)	100%	Per Applicable Device Specification (Note 6)	—
11. Burn-In Test	1015 240 Hrs. @ 125°C Min. (Cond. F Not Allowed)	100%	1015 160 Hrs. @ 125°C Min.	100%
12. Interim (Post-Burn-In) Electrical Parameters	Per Applicable Device Specification (Note 13)	100%		
13. Reverse Bias Burn-In (Note 7)	1015; Test Condition A, C, 72 Hrs. @ 150°C Min. (Cond. F Not Allowed)	100%		—
15. PDA Calculation	5% Parametric (Note 14), 3% Functional −25°C	All Lots	5% Parametric (Note 14)	All Lots
16. Final Electrical Test	Per Applicable Device Specification		Per Applicable Device Specification	
a) Static Tests				
1) 25°C (Subgroup 1, Table I, 5005)		100%		100%
2) Max & Min Rated Operating Temp. (Subgroups 2, 3, Table I, 5005)		100%		100%
b) Dynamic Tests & Switching Tests, 25°C (Subgroups 4, 9, Table I, 5005)		100%		100%
c) Functional Test, 25°C (Subgroup 7, Table I, 5005)		100%		100%

TABLE III. 100% Screening Requirements (Continued)

Screen	Class S		Class B	
	Method	Reqmt	Method	Reqmt
17. Seal Fine, Gross	1014	100%, (Note 8)	1014	100%, (Note 9)
18. Radiographic (Note 10)	2012 Two Views	100%		—
19. Qualification or Quality Conformance Inspection Test Sample Selection	(Note 11)	Samp.	(Note 11)	Samp.
20. External Visual (Note 12)	2009	100%		100%

Note 1: Unless otherwise specified, at the manufacturer's option, test samples for Group B, bond strength (Method 5005) may be randomly selected prior to or following internal visual (Method 5004), prior to sealing provided all other specification requirements are satisfied (e.g. bond strength requirements shall apply to each inspection lot, bond failures shall be counted even if the bond would have failed internal visual).

Note 2: For Class B devices, this test may be replaced with thermal shock method 1011, test condition A, minimum.

Note 3: At the manufacturer's option, visual inspection for catastrophic failures may be conducted after each of the thermal/mechanical screens, after the sequence or after seal test. Catastrophic failures are defined as missing leads, broken packages, or lids off.

Note 4: The PIND test may be performed in any sequence after step 6 and prior to step 16. See MIL-M-38510, paragraph 4.6.3.

Note 5: Class S devices shall be serialized prior to interim electrical parameter measurements.

Note 6: When specified, all devices shall be tested for those parameters requiring delta calculations.

Note 7: Reverse bias burn-in is a requirement only when specified in the applicable device specification. The order of performing burn-in and reverse bias burn-in may be inverted.

Note 8: For Class S devices, the seal test may be performed in any sequence between step 16 and step 19, but it shall be performed after all shearing and forming operations on the terminals.

Note 9: For Class B devices, the fine and gross seal tests shall be performed separate or together in any sequence and order between step 6 and step 20 except that they shall be performed after all shearing and forming operations on the terminals. When 100% seal screen cannot be performed after shearing and forming (e.g. flatpacks and chip carriers) the seal screen shall be done 100% prior to these operations and a sample test (LTPD = 5) shall be performed on each inspection lot following these operations. If the sample fails, 100% rescreening shall be required.

Note 10: The radiographic screen may be performed in any sequence after step 19.

Note 11: Samples shall be selected for testing in accordance with the specific device class and lot requirements of Method 5005

Note 12: External Visual shall be performed on the lot any time after step 19 and prior to shipment.

Note 13: Read and Record when past burn-in delta measurements are specified.

Note 14: PDA shall apply to all static, dynamic, functional, and switching measurements at either 25°C or maximum rated operating temperature.

Military Analog Products Available From National Semiconductor

Listed below are the military class B Analog devices available from National Semiconductor. Many of these are also available as Class S product. Additional information including new product plans can be obtained from our sales offices.

Device Type	Mil * Class B	883 Class B	Desc	JAN
AH0014D	x			
AH0015D	x			
AH0019D	x			
LF111H	x			
LF11201D		x		
LF11202D		x		
LF11331D		x		
LF11332D		x		
LF11333D		x		
LF11508D	x			
LF11509D	x			
LF147D		x		
LF155AH		x		
LF155H		x		x
LF155J-8				x
LF155W				x
LF156AH		x		
LF156H		x		x
LF156J-8				x
LF156W				x
LF157AH		x		
LF157H		x		
LF198H		x		
LF411MH		x		x
LF411W				x
LF412MH		x		x
LF441MH	x			
LF442MH		x		
LF444MD		x		
LH0002H		x	x	
LH0003H	x			
LH0004H	x			
LH0020G	x			
LH0021K	x			
LH0022D	x			
LH0022H	x			
LH0023G	x			
LH0024H	x			

Device Type	Mil * Class B	883 Class B	Desc	JAN
LH0032G	x		x	
LH0033AG	x			
LH0033G	x		x	
LH0036G	x			
LH0038D	x			
LH0041G	x			
LH0042D	x			
LH0042H	x			
LH0043G	x			
LH0044AH	x			
LH0044H	x			
LH0052H	x			
LH0053G	x			
LH0061K	x			
LH0062D	x			
LH0062H	x			
LH0063K	x			
LH0070-0H	x			
LH0070-1H	x			
LH0070-2H	x			
LH0071-0H	x			
LH0071-1H	x			
LH0071-2H	x			
LH0075G	x			
LH0076G	x			
LH0082D	x			
LH0084D	x			
LH0086D	x			
LH0091D	x			
LH0094D	x			
LH00101AK	x			
LH0101K	x			
LH2101AD		x		
LH2108AD		x		
LH2108D		x		
LH2110D		x		
LH2111D		x		
LH2111F	x			

*Some older products are not completely compliant with MIL-STD-883 but are still required for use in military systems. These devices are screened to the same stringent requirements as 883 product but are marked "-MIL".

Military Analog Products Available From National Semiconductor

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Device Type	Mil * Class B	883 Class B	Desc	JAN
LH24250F	x			
LM10H		x		
LM101AH		x		x
LM101AJ-14		x		x
LM101AJ		x		
LM101AW				x
LM102H		x		
LM103H-3.0		x	x	
LM103H-3.3		x	x	
LM103H-3.6		x	x	
LM103H-3.9		x	x	
LM104H		x		
LM105H		x		
LM106H		x		
LM107H		x		
LM107J-14		x		
LM107J		x		
LM108AH		x		x
LM108AJ-8		x		x
LM108AJ		x		
LM108AW				x
LM108H		x		
LM108J-8		x		
LM108J		x		
LM109H		x		
LM109KSTEEL		x		
LM11H		x		
LM110H		x		
LM110J-8		x		
LM110J		x		
LM111H		x		x
LM111J		x		x
LM111W				x
LM112H		x		
LM113-1H		x	x	
LM113-2H		x	x	
LM113H		x	x	
LM117H		x	x	x

Device Type	Mil * Class B	883 Class B	Desc	JAN
LM117HVH		x	x	
LM117HVKSTL		x	x	
LM117KSTEEL		x	x	x
LM118H		x		x
LM118J-8		x		x
LM118J		x		
LM118W				x
LM119H		x	x	
LM119J		x	x	
LM120H-12		x		
LM120H-15		x		
LM120H-5.0		x		
LM120K-12		x		
LM120K-15		x		
LM120K-5.0		x		
LM121AH		x		
LM121H		x		
LM122H		x		
LM123KSTEEL		x		
LM124AJ		x		
LM124J		x		x
LM125H		x		
LM126H		x		
LM129AH		x		
LM129BH		x		
LM131AH		x		
LM131H		x		
LM135H		x		
LM136AH-2.5		x	x	
LM136H-2.5		x		
LM136H-5.0		x		
LM137H		x	x	
LM137HVH		x	x	
LM137HVKSTEEL		x	x	
LM137KSTEEL		x	x	
LM138KSTEEL		x		
LM139AJ		x		
LM139J		x		x

*Some older products are not completely compliant with MIL-STD-883 but are still required for use in military systems. These devices are screened to the same stringent requirements as 883 product but are marked "-MIL".

Military Analog Products Available From National Semiconductor

Listed below are the military class B Analog devices available from National Semiconductor. Many of these are also available as Class S product. Additional information including new product plans can be obtained from our sales offices.

Device Type	Mil * Class B	883 Class B	Desc	JAN
LM139W				x
LM140AK-12		x		
LM140AK-15		x		
LM140AK-5.0		x		
LM140K-12		x		
LM140K-15		x		
LM140K-5.0		x		
LM140LAH-12		x		
LM140LAH-15		x		
LM140LAH-5.0		x		
LM143H		x	x	
LM144H		x	x	
LM145K-5.0		x		
LM145K-5.2		x		
LM146J		x		
LM148J		x		x
LM149J		x		
LM150KSTEEL	x			
LM1536H		x	x	
LM1558H		x		
LM1558J		x		
LM158AH		x		
LM158AJ		x		
LM158H		x		
LM158J		x		
LM1596H	x			
LM160H		x		
LM160J-14		x		
LM160J		x		
LM161F	x			
LM161H		x		
LM161J		x		
LM185BXH-1.2		x		
LM185BYH-1.2		x		

Device Type	Mil * Class B	883 Class B	Desc	JAN
LM185H-1.2		x		
LM193AH		x		
LM193H		x		x
LM193J-8				x
LM193W				x
LM194H		x		
LM195H		x		
LM195K		x		
LM199AH-20		x		
LM199AH		x		
LM199H		x		
LM4250H	x			
LM4250J	x			
LM555H		x		
LM555J		x		
LM556J	x			
LM567H		x		
LM709AH		x		
LM709H		x		
LM710H		x		
LM723H		x		
LM723J				x
LM725H		x		
LM733H	x			
LM741AJ-14		x		
LM741AJ		x		
LM741H		x		x
LM7415-14		x		
LM741J		x		x
LM741W				x
LM747H		x		x
LM747J		x		
LM748H		x		
LM748J		x		

*Some older products are not completely compliant with MIL-STD-883 but are still required for use in military systems. These devices are screened to the same stringent requirements as 883 product but are marked "-MIL".

National Semiconductor

Appendix E
Understanding Integrated Circuit Package Power Capabilities

INTRODUCTION

The short and long term reliability of National Semiconductor's interface circuits, like any integrated circuit, is very dependent on its environmental condition. Beyond the mechanical/environmental factors, nothing has a greater influence on this reliability than the electrical and thermal stress seen by the integrated circuit. Both of these stress issues are specifically addressed on every interface circuit data sheet, under the headings of Absolute Maximum Ratings and Recommended Operating Conditions.

However, through application calls, it has become clear that electrical stress conditions are generally more understood than the thermal stress conditions. Understanding the importance of electrical stress should never be reduced, but clearly, a higher focus and understanding must be placed on thermal stress. Thermal stress and its application to interface circuits from National Semiconductor is the subject of this application note.

FACTORS AFFECTING DEVICE RELIABILITY

Figure 1 shows the well known "bathtub" curve plotting failure rate versus time. Similar to all system hardware (mechanical or electrical) the reliability of interface integrated circuits conform to this curve. The key issues associated with this curve are infant mortality, failure rate, and useful life.

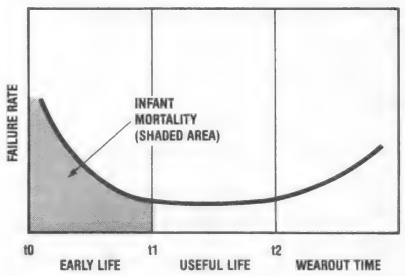

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FIGURE 1. Failure Rate vs Time

Infant mortality, the high failure rate from time t0 to t1 (early life), is greatly influenced by system stress conditions other than temperature, and can vary widely from one application to another. The main stress factors that contribute to infant mortality are electrical transients and noise, mechanical maltreatment and excessive temperatures. Most of these failures are discovered in device test, burn-in, card assembly and handling, and initial system test and operation. Although important, much literature is available on the subject of infant mortality in integrated circuits and is beyond the scope of this application note.

Failure rate is the number of devices that will be expected to fail in a given period of time (such as, per million hours). The mean time between failure (MTBF) is the average time (in hours) that will be expected to elapse after a unit has failed before the next unit failure will occur. These two primary "units of measure" for device reliability are inversely related:

$$\text{MTBF} = \frac{1}{\text{Failure Rate}}$$

Although the "bathtub" curve plots the overall failure rate versus time, the useful failure rate can be defined as the percentage of devices that fail per-unit-time during the flat portion of the curve. This area, called the useful life, extends between t1 and t2 or from the end of infant mortality to the onset of wearout. The useful life may be as short as several years but usually extends for decades if adequate design margins are used in the development of a system.

Many factors influence useful life including: pressure, mechanical stress, thermal cycling, and electrical stress. However, die temperature during the device's useful life plays an equally important role in triggering the onset of wearout.

FAILURE RATES vs TIME AND TEMPERATURE

The relationship between integrated circuit failure rates and time and temperature is a well established fact. The occurrence of these failures is a function which can be represented by the Arrhenius Model. Well validated and predominantly used for accelerated life testing of integrated circuits, the Arrhenius Model assumes the degradation of a performance parameter is linear with time and that MTBF is a function of temperature stress. The temperature dependence is an exponential function that defines the probability of occurrence. This results in a formula for expressing the lifetime or MTBF at a given temperature stress in relation to another MTBF at a different temperature. The ratio of these two MTBFs is called the acceleration factor F and is defined by the following equation:

$$F = \frac{X1}{X2} = \exp\left[\frac{E}{K}\left(\frac{1}{T2} - \frac{1}{T1}\right)\right]$$

Where: X1 = Failure rate at junction temperature T1
X2 = Failure rate at junction temperature T2
T = Junction temperature in degrees Kelvin
E = Thermal activation energy in electron volts (ev)
K = Boltzman's constant

However, the dramatic acceleration effect of junction temperature (chip temperature) on failure rate is illustrated in a plot of the above equation for three different activation energies in *Figure 2*. This graph clearly demonstrates the importance of the relationship of junction temperature to device failure rate. For example, using the 0.99 ev line, a 30° rise in junction temperature, say from 130°C to 160°C, results in a 10 to 1 increase in failure rate.

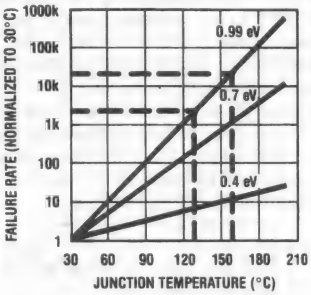

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FIGURE 2. Failure Rate as a Function of Junction Temperature

DEVICE THERMAL CAPABILITIES

There are many factors which affect the thermal capability of an integrated circuit. To understand these we need to understand the predominant paths for heat to transfer out of the integrated circuit package. This is illustrated by *Figures 3* and *4*.

Figure 3 shows a cross-sectional view of an assembled integrated circuit mounted into a printed circuit board.

Figure 4 is a flow chart showing how the heat generated at the power source, the junctions of the integrated circuit flows from the chip to the ultimate heat sink, the ambient environment. There are two predominant paths. The first is from the die to the die attach pad to the surrounding package material to the package lead frame to the printed circuit board and then to the ambient. The second path is from the package directly to the ambient air.

Improving the thermal characteristics of any stage in the flow chart of *Figure 4* will result in an improvement in device thermal characteristics. However, grouping all these characteristics into one equation determining the overall thermal capability of an integrated circuit/package/environmental condition is possible. The equation that expresses this relationship is:

$$T_J = T_A + P_D (\theta_{JA})$$

Where: T_J = Die junction temperature

T_A = Ambient temperature in the vicinity device

P_D = Total power dissipation (in watts)

θ_{JA} = Thermal resistance junction-to-ambient

θ_{JA}, the thermal resistance from device junction-to-ambient temperature, is measured and specified by the manufacturers of integrated circuits. National Semiconductor utilizes special vehicles and methods to measure and monitor this parameter. All circuit data sheets specify the thermal characteristics and capabilities of the packages available for a given device under specific conditions—these package power ratings directly relate to thermal resistance junction-to-ambient or θ_{JA}.

Although National provides these thermal ratings, it is critical that the end user understand how to use these numbers to improve thermal characteristics in the development of his system using IC components.

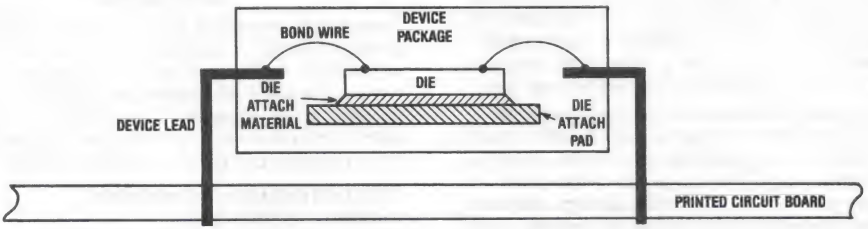

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FIGURE 3. Integrated Circuit Soldered into a Printed Circuit Board (Cross-Sectional View)

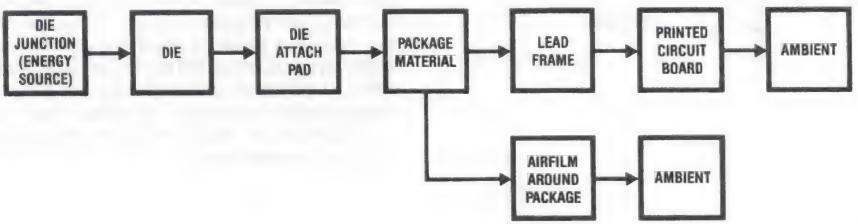

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FIGURE 4. Thermal Flow (Predominant Paths)

DETERMINING DEVICE OPERATING JUNCTION TEMPERATURE

From the above equation the method of determining actual worst-case device operating junction temperature becomes straightforward. Given a package thermal characteristic, θ_{JA}, worst-case ambient operating temperature, T_A(max), the only unknown parameter is device power dissipation, P_D. In calculating this parameter, the dissipation of the integrated circuit due to its own supply has to be considered, the dissipation within the package due to the external load must also be added. The power associated with the load in a dynamic (switching) situation must also be considered. For example, the power associated with an inductor or a capacitor in a static versus dynamic (say, 1 MHz) condition is significantly different.

The junction temperature of a device with a total package power of 600 mW at 70°C in a package with a thermal resistance of 63°C/W is 108°C.

$$T_J = 70°C + (63°C/W) \times (0.6W) = 108°C$$

The next obvious question is, "how safe is 108°C?"

MAXIMUM ALLOWABLE JUNCTION TEMPERATURES

What is an acceptable maximum operating junction temperature is in itself somewhat of a difficult question to answer. Many companies have established their own standards based on corporate policy. However, the semiconductor industry has developed some defacto standards based on the device package type. These have been well accepted as numbers that relate to reasonable (acceptable) device lifetimes, thus failure rates.

National Semiconductor has adopted these industry-wide standards. For devices fabricated in a molded package, the maximum allowable junction temperature is 150°C. For these devices assembled in ceramic or cavity DIP packages, the maximum allowable junction temperature is 175°C. The numbers are different because of the differences in package types. The thermal strain associated with the die package interface in a cavity package is much less than that exhibited in a molded package where the integrated circuit chip is in direct contact with the package material.

Let us use this new information and our thermal equation to construct a graph which displays the safe thermal (power) operating area for a given package type. *Figure 5* is an example of such a graph. The end points of this graph are easily determined. For a 16-pin molded package, the maximum allowable temperature is 150°C; at this point no power dissipation is allowable. The power capability at 25°C is 1.98W as given by the following calculation:

$$P_D @ 25°C = \frac{T_J(max) - T_A}{\theta_{JA}} = \frac{150°C - 25°C}{63°C/W} = 1.98W$$

The slope of the straight line between these two points is minus the inversion of the thermal resistance. This is referred to as the derating factor.

$$\text{Derating Factor} = -\frac{1}{\theta_{JA}}$$

As mentioned, *Figure 5* is a plot of the safe thermal operating area for a device in a 16-pin molded DIP. As long as the intersection of a vertical line defining the maximum ambient temperature (70°C in our previous example) and maximum device package power (600 mW) remains below the maximum package thermal capability line the junction temperature will remain below 150°C—the limit for a molded package. If the intersection of ambient temperature and package power fails on this line, the maximum junction temperature will be 150°C. Any intersection that occurs above this line will result in a junction temperature in excess of 150°C and is not an appropriate operating condition.

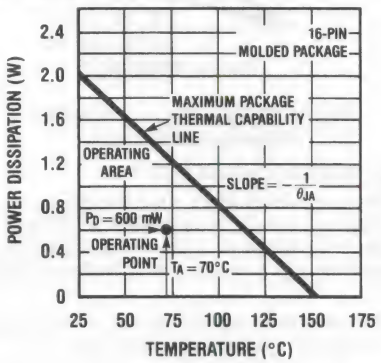

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FIGURE 5. Package Power Capability vs Temperature

The thermal capabilities of all integrated circuits are expressed as a power capability at 25°C still air environment with a given derating factor. This simply states, for every degree of ambient temperature rise above 25°C, reduce the package power capability stated by the derating factor which is expressed in mW/°C. For our example—a θ_{JA} of 63°C/W relates to a derating factor of 15.9 mW/°C.

FACTORS INFLUENCING PACKAGE THERMAL RESISTANCE

As discussed earlier, improving any portion of the two primary thermal flow paths will result in an improvement in overall thermal resistance junction-to-ambient. This section discusses those components of thermal resistance that can be influenced by the manufacturer of the integrated circuit. It also discusses those factors in the overall thermal resistance that can be impacted by the end user of the integrated circuit. Understanding these issues will go a long way in understanding chip power capabilities and what can be done to insure the best possible operating conditions and, thus, best overall reliability.

Die Size

Figure 6 shows a graph of our 16-pin DIP thermal resistance as a function of integrated circuit die size. Clearly, as the chip size increases the thermal resistance decreases—this relates directly to having a larger area with which to dissipate a given power.

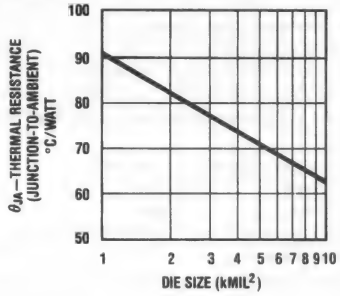

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FIGURE 6. Thermal Resistance vs Die Size

Lead Frame Material

Figure 7 shows the influence of lead frame material (both die attach and device pins) on thermal resistance. This graph compares our same 16-pin DIP with a copper lead frame, a Kovar lead frame, and finally an Alloy 42 type lead frame—these are lead frame materials commonly used in the industry. Obviously the thermal conductivity of the lead frame material has a significant impact in package power capability. Molded interface circuits from National Semiconductor use the copper lead frame exclusively.

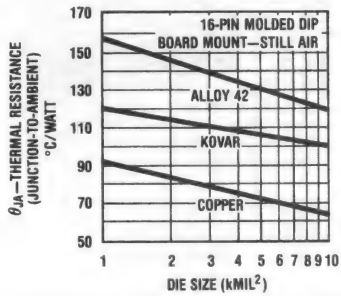

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FIGURE 7. Thermal Resistance vs Lead Frame Material

Board vs Socket Mount

One of the major paths of dissipating energy generated by the integrated circuit is through the device leads. As a result of this, the graph of *Figure 8* comes as no surprise. This compares the thermal resistance of our 16-pin package soldered into a printed circuit board (board mount) compared to the same package placed in a socket (socket mount). Adding a socket in the path between the PC board and the device adds another stage in the thermal flow path, thus increasing the overall thermal resistance. The thermal capabilities of National Semiconductor's interface circuits are specified assuming board mount conditions. If the devices are placed in a socket the thermal capabilities should be reduced by approximately 5% to 10%.

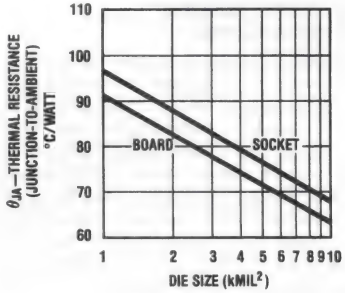

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FIGURE 8. Thermal Resistance vs Board or Socket Mount

Air Flow

When a high power situation exists and the ambient temperature cannot be reduced, the next best thing is to provide air flow in the vicinity of the package. The graph of *Figure 9* illustrates the impact this has on thermal resistance. This graph plots the relative reduction in thermal resistance normalized to the still air condition for our 16-pin molded DIP. The thermal ratings on National Semiconductor's interface circuits data sheets relate to the still air environment.

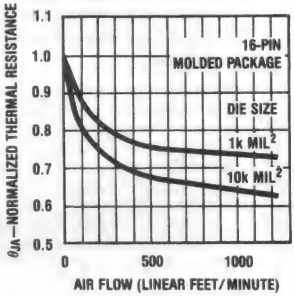

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FIGURE 9. Thermal Resistance vs Air Flow

Other Factors

A number of other factors influence thermal resistance. The most important of these is using thermal epoxy in mounting ICs to the PC board and heat sinks. Generally these techniques are required only in the very highest of power applications.

Some confusion exists between the difference in thermal resistance junction-to-ambient (θ_{JA}) and thermal resistance junction-to-case (θ_{JC}). The best measure of actual junction temperature is the junction-to-ambient number since nearly all systems operate in an open air environment. The only situation where thermal resistance junction-to-case is important is when the entire system is immersed in a thermal bath and the environmental temperature is indeed the case temperature. This is only used in extreme cases and is the exception to the rule and, for this reason, is not addressed in this application note.

NATIONAL SEMICONDUCTOR PACKAGE CAPABILITIES

Figures 10 and *11* show composite plots of the thermal characteristics of the most common package types in the National Semiconductor Linear Circuits product family. *Figure 10* is a composite of the copper lead frame molded package. *Figure 11* is a composite of the ceramic (cavity) DIP using poly die attach. These graphs represent board mount still air thermal capabilities. Another, and final, thermal resistance trend will be noticed in these graphs. As the number of device pins increase in a DIP the thermal resistance decreases. Referring back to the thermal flow chart, this trend should, by now, be obvious.

RATINGS ON INTERFACE CIRCUITS DATA SHEETS

In conclusion, all National Semiconductor Linear Products define power dissipation (thermal) capability. This information can be found in the Absolute Maximum Ratings section of the data sheet. The thermal information shown in this application note represents average data for characterization of the indicated package. Actual thermal resistance can vary from ±10% to ±15% due to fluctuations in assembly quality, die shape, die thickness, distribution of heat sources on the die, etc. The numbers quoted in the linear data sheets reflect a 15% safety margin from the average numbers found in this application note. Insuring that total package power remains under a specified level will guarantee that the maximum junction temperature will not exceed the package maximum.

The package power ratings are specified as a maximum power at 25°C ambient with an associated derating factor for ambient temperatures above 25°C. It is easy to determine the power capability at an elevated temperature. The power specified at 25°C should be reduced by the derating factor for every degree of ambient temperature above 25°C. For example, in a given product data sheet the following will be found:

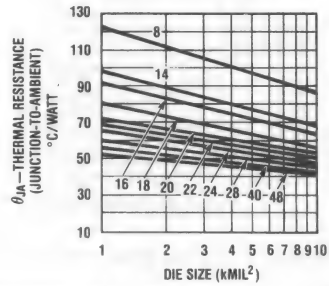
Maximum Power Dissipation* at 25°C
Cavity Package 1509 mW
Molded Package 1476 mW

* Derate cavity package at 10 mW/°C above 25°C; derate molded package at 11.8 mW/°C above 25°C.

If the molded package is used at a maximum ambient temperature of 70°C, the package power capability is 945 mW.

$$P_D @ 70°C = 1476\ mW - (11.8\ mW/°C) \times (70°C - 25°C)$$
$$= 945\ mW$$

Molded (N Package) DIP* Copper Leadframe—HTP Die Attach Board Mount—Still Air



*Packages from 8- to 20-pin 0.3 mil width
22-pin 0.4 mil width
24- to 40-pin 0.6 mil width

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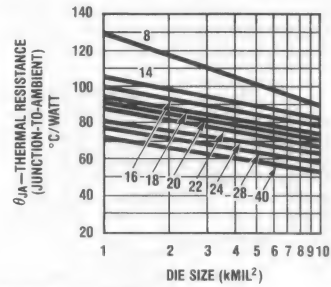
FIGURE 10. Thermal Resistance vs Die Size vs Package Type (Molded Package)

Cavity (J Package) DIP* Poly Die Attach Board Mount—Still Air



*Packages from 8- to 20-pin 0.3 mil width
22-pin 0.4 mil width
24- to 48-pin 0.6 mil width

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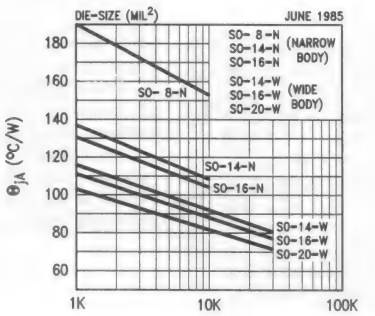
FIGURE 11. Thermal Resistance vs Die Size vs Package Type (Cavity Package)



Θ_{JA} – THERMAL RESISTANCE FOR "SO" PACKAGES (BOARD MOUNT)

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FIGURE 12

National Semiconductor

APPENDIX F
How to Get the Right Information From a Data Sheet

Not All Data Sheets Are Created Alike, and False Assumptions Could Cost an Engineer Time and Money

By Robert A. Pease

When a new product arrives in the marketplace, it hopefully will have a good, clear data sheet with it.

The data sheet can show the prospective user how to apply the device, what performance specifications are guaranteed and various typical applications and characteristics. If the data-sheet writer has done a good job, the user can decide if the product will be valuable to him, exactly how well it will be of use to him and what precautions to take to avoid problems.

SPECIFICATIONS

The most important area of a data sheet specifies the characteristics that are guaranteed—and the test conditions that apply when the tests are done. Ideally, all specifications that the users will need will be spelled out clearly. If the product is similar to existing products, one can expect the data sheet to have a format similar to other devices.

But, if there are significant changes and improvements that nobody has seen before, then the writer must clarify what is meant by each specification. Definitions of new phrases or characteristics may even have to be added as an appendix.

For example, when fast-settling operational amplifiers were first introduced, some manufacturers defined settling time as the time after slewing before the output finally enters and stays within the error-band; but other manufacturers included the slewing time in their definition. Because both groups made their definitions clear, the user was unlikely to be confused or misled.

However, the reader ought to be on the alert. In a few cases, the data-sheet writer is playing a specsmanship game, and is trying to show an inferior (to some users) aspect of a product in a light that makes it look superior (which it may be, to a couple of users).

GUARANTEES

When a data sheet specifies a guaranteed minimum value, what does it mean? An assumption might be made that the manufacturer has actually tested that specification and has great confidence that no part could fail that test and still be shipped. Yet that is not always the case.

For instance, in the early days of op amps (20 years ago), the differential-input impedance might have been guaranteed at 1 MΩ—but the manufacturer obviously did not measure the impedance. When a customer insisted, "I have to know how you measure this impedance," it had to be explained that the impedance was not measured, but that the base current was. The correlation between I_b and Z_{in} permitted the substitution of this simple dc test for a rather messy, noisy, hard-to-interpret test.

Every year, for the last 20 years, manufacturers have been trying to explain, with varying success, why they do not measure the Z_{in} *per se,* even though they do guarantee it.

In other cases, the manufacturer may specify a test that can be made only on the die as it is probed on the wafer, but cannot be tested after the die is packaged because that signal is not accessible any longer. To avoid frustrating and confusing the customer, some manufacturers are establishing two classes of guaranteed specifications:

- The tested limit represents a test that cannot be doubted, one that is actually performed directly on 100 percent of the devices, 100 percent of the time.
- The design limit covers other tests that may be indirect, implicit or simply guaranteed by the inherent design of the device, and is unlikely to cause a failure rate (on that test), even as high as one part per thousand.

Why was this distinction made? Not just because customers wanted to know which specifications were guaranteed by testing, but because the quality-assurance group insisted that it was essential to separate the tested guarantees from the design limits so that the AQL (assurance-quality level) could be improved from 0.1 percent to down below 100 ppm.

Some data sheets guarantee characteristics that are quite expensive and difficult to test (even harder than noise) such as long-term drift (20 ppm or 50 ppm over 1,000 hours).

The data sheet may not tell the reader if it is measured, tested or estimated. One manufacturer may perform a 100-percent test, while another states, "Guaranteed by sample testing." This is not a very comforting assurance that a part is good, especially in a critical case where only a long-term test can prove if the device did meet the manufacturer's specification. If in doubt, question the manufacturer.

TYPICALS

Next to a guaranteed specification, there is likely to be another in a column labeled "typical".

It might mean that the manufacturer once actually saw one part as good as that. It could indicate that half the parts are better than that specification, and half will be worse. But it is equally likely to mean that, five years ago, half the parts were better and half worse. It could easily signify that a few parts might be slightly better, and a few parts a lot worse; after all, if the noise of an amplifier is extremely close to the theoretical limit, one cannot expect to find anything much better than that, but there will always be a few noisy ones.

If the specification of interest happens to be the bias current (I_b) of an op amp, a user can expect broad variations. For example, if the specification is 200 nA maximum, there might be many parts where I_b is 40 nA on one batch (where the beta is high), and a month later, many parts where the I_b is 140 nA when the beta is low.

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Absolute Maximum Ratings (Note 11)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage	+35V to −0.2V
Output Voltage	+6V to −1.0V
Output Current	10 mA
Storage Temperature,	
TO-46 Package	−76°F to +356°F
TO-92 Package	−76°F to +300°F
Lead Temp. (Soldering, 4 seconds) *	
TO-46 Package	+300°C
TO-92 Package	+260°C
Specified Operating Temp. Range (Note 2)	
	T_{MIN} to T_{MAX}
LM34, LM34A	−50°F to +300°F
LM34C, LM34CA	−40°F to +230°F
LM34D	+32°F to +212°F

DC Electrical Characteristics (Note 1, **Note 6**)

Parameter	Conditions	LM34A			LM34CA			Units (Max)
		Typical	Tested Limit (Note 4)	Design Limit (Note 5)	Typical	Tested Limit (Note 4)	Design Limit (Note 5)	
Accuracy (Note 7)	T_A = +77°F	±0.4	±1.0		±0.4	±1.0		°F
	T_A = 0°F	±0.6			±0.6		±2.0	°F
	T_A = T_{MAX}	±0.8	±2.0		±0.8	±2.0		°F
	T_A = T_{MIN}	±0.8	±2.0		±0.8		±3.0	°F
Nonlinearity (Note 8)	$T_{MIN} \le T_A \le T_{MAX}$	**±0.35**		**±0.7**	**±0.30**		**±0.6**	°F
Sensor Gain (Average Slope)	$T_{MIN} \le T_A \le T_{MAX}$	**+10.0**	**+9.9, +10.1**		**+10.0**		**+9.9, +10.1**	mV/°F, min mV/°F, max
Load Regulation (Note 3)	T_A = +77°F	±0.4	±1.0		±0.4	±1.0		mV/mA
	$T_{MIN} \le T_A \le T_{MAX}$	**±0.5**		**±3.0**	**±0.5**		**±3.0**	mV/mA
	$0 \le I_L \le 1$ mA							
Line Regulation (Note 3)	T_A = +77°F	±0.01	±0.05		±0.01	±0.05		mV/V
	$5V \le V_S \le 30V$	**±0.02**		**±0.1**	**±0.02**		**±0.1**	mV/V
Quiescent Current (Note 9)	V_S = +5V, +77°F	75	90		75	90		µA
	V_S = +5V	**131**		**160**	**116**		**139**	µA
	V_S = +30V, +77°F	76	92		76	92		µA
	V_S = +30V	**132**		**163**	**117**		**142**	µA
Change of Quiescent Current (Note 3)	$4V \le V_S \le 30V$, +77°F	+0.5	2.0		0.5	2.0		µA
	$5V \le V_S \le 30V$	**+1.0**		**3.0**	**1.0**		**3.0**	µA
Temperature Coefficient of Quiescent Current		**+0.30**		**+0.5**	**+0.30**		**+0.5**	µA/°F
Minimum Temperature for Rated Accuracy	In circuit of *Figure 1*, I_L = 0	+3.0		+5.0	+3.0		+5.0	°F
Long-Term Stability	$T_j = T_{MAX}$ for 1000 hours	±0.16			±0.16			°F

Note 1: Unless otherwise noted, these specifications apply: $-50°F \le T_j \le +300°F$ for the LM34 and LM34A; $-40°F \le T_j \le +230°F$ for the LM34C and LM34CA; and $+32°F \le T_j \le +212°F$ for the LM34D. V_S = +5 Vdc and I_{LOAD} = 50 µA in the circuit of *Figure 2;* +6 Vdc for LM34 and LM34A for $230°F \le T_j \le 300°F$. These specifications also apply from +5°F to T_{MAX} in the circuit of *Figure 1.*

Note 2: Thermal resistance of the TO-46 package is 292°F/W junction to ambient and 43°F/W junction to case. Thermal resistance of the TO-92 package is 324°F/W junction to ambient.

Note 3: Regulation is measured at constant junction temperature using pulse testing with a low duty cycle. Changes in output due to heating effects can be computed by multiplying the internal dissipation by the thermal resistance.

Note 4: Tested limits are guaranteed and 100% tested in production.

Note 5: Design limits are guaranteed (but not 100% production tested) over the indicated temperature and supply voltage ranges. These limits are not used to calculate outgoing quality levels.

Note 6: Specification in **BOLDFACE TYPE** apply over the full rated temperature range.

Note 7: Accuracy is defined as the error between the output voltage and 10 mV/°F times the device's case temperature at specified conditions of voltage, current, and temperature (expressed in °F).

Note 8: Nonlinearity is defined as the deviation of the output-voltage-versus-temperature curve from the best-fit straight line over the device's rated temperature range.

Note 9: Quiescent current is defined in the circuit of *Figure 1.*

Note 10: Contact factory for availability of LM34CAZ.

** **Note 11:** Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. DC and AC electrical specifications do not apply when operating the device beyond its rated operating conditions (see Note 1).

A Point-By-Point Look

Let's look a little more closely at the data sheet of the National Semiconductor LM34, which happens to be a temperature sensor.

Note 1 lists the nominal test conditions and test circuits in which all the characteristics are defined. Some additional test conditions are listed in the column "Conditions", but Note 1 helps minimize the clutter.

Note 2 gives the thermal impedance, (which may also be shown in a chart or table).

Note 3 warns that an output impedance test, if done with a long pulse, could cause significant self-heating and thus, error.

Note 6 is intended to show which specs apply at all rated temperatures.

Note 7 is the definition of the "Accuracy" spec, and Note 8 the definition for non-linearity. Note 9 states in what test circuit the quiescent current is defined. Note 10 indicates that one model of the family may not be available at the time of printing (but happens to be available now), and Note 11 is the definition of Absolute Max Ratings.

* Note—the "4 seconds" soldering time is a new standard for plastic packages.

** Note—the wording of Note 11 has been revised—this is the best wording we can devise, and we will use it on all future datasheets.

APPLICATIONS

Another important part of the data sheet is the applications section. It indicates the novel and conventional ways to use a device. Sometimes these applications are just little ideas to tweak a reader's mind. After looking at a couple of applications, one can invent other ideas that are useful. Some applications may be of no real interest or use.

In other cases, an application circuit may be the complete definition of the system's performance; it can be the test circuit in which the specification limits are defined, tested and guaranteed. But, in all other instances, the performance of a typical application circuit is not guaranteed, it is only typical. In many circumstances, the performance may depend on external components and their precision and matching. Some manufacturers have added a phrase to their data sheets:

"Applications for any circuits contained in this document are for illustration purposes only and the manufacturer makes no representation or warranty that such applications will be suitable for the use indicated without further testing or modification."

In the future, manufacturers may find it necessary to add disclaimers of this kind to avoid disappointing users with circuits that work well, much of the time, but cannot be easily guaranteed.

The applications section is also a good place to look for advice on quirks—potential drawbacks or little details that may not be so little when a user wants to know if a device will actually deliver the expected performance.

For example, if a buffer can drive heavy loads and can handle fast signals cleanly (at no load), the maker isn't doing anybody any favors if there is no mention that the distortion goes sky-high if the rated load is applied.

Another example is the application hint for the LF156 family:

"Exceeding the negative common-mode limit on either input will cause a reversal of the phase to output and force the amplifier output to the corresponding high or low state. Exceeding the negative common-mode limit on both inputs will force the amplifier output to a high state. In neither case does a latch occur, since raising the input back within the common-mode range again puts the input stage and, thus the amplifier, in a normal operating mode."

That's the kind of information a manufacturer should really give to a data-sheet reader because no one could ever guess it.

Sometimes, a writer slips a quirk into a characteristic curve, but it's wiser to draw attention to it with a line of text. This is because it's better to make the user sad before one gets started, rather than when one goes into production. Conversely, if a user is going to spend more than 10 minutes using a new product, one ought to spend a full five minutes reading the entire data sheet.

FINE PRINT

What other fine print can be found on a data sheet? Sometimes the front page may be marked "advance" or "preliminary." Then on the back page, the fine print may say something such as:

"This data sheet contains preliminary limits and design specifications. Supplemental information will be published at a later date. The manufacturer reserves the right to make changes in the products contained in this document in order to improve design or performance and to supply the best possible products. We also assume no responsibility for the use of any circuits described herein, convey no license under any patent or other right and make no representation that the circuits are free from patent infringement."

In fact, after a device is released to the marketplace in a preliminary status, the engineers love to make small improvements and upgrades in specifications and characteristics, and hate to degrade a specification from its first published value—but occasionally that is necessary.

Another item in the fine print is the manufacturer's telephone number. Usually it is best to refer questions to the local sales representative or field-applications engineer, because they may know the answer or they may be best able to put a questioner in touch with the right person at the factory.

Occasionally, the factory's applications engineers have all the information. Other times, they have to bring in product engineers, test engineers or marketing people. And sometimes the answer can't be generated quickly—data have to be gathered, opinions solidified or policies formulated before the manufacturer can answer the question. Still, the telephone number is the key to getting the factory to help.

ORIGINS OF DATA SHEETS

Of course, historically, most data sheets for a class of products have been closely modeled on the data sheet of the forerunner of that class. The first data sheet was copied to make new versions.

That's the way it happened with the UA709 (the first monolithic op amp) and all its copies, as well as many other similar families of circuits.

Even today, an attempt is made to build on the good things learned from the past and add a few improvements when necessary. But, it's important to have real improvements, not just change for the sake of change.

So, while it's not easy to get the format and everything in it exactly right to please everybody, new data sheets are continually surfacing with new features, applications ideas, specifications and aids for the user. And, if the users complain loudly enough about misleading or inadequate data sheets, they can help lead the way to change data sheets. That's how many of today's improvements came about—through customer demand.

Who writes data sheets? In some cases, a marketing person does the actual writing and engineers do the checking. In other companies, the engineer writes, while marketing people and other engineers check. Sometimes, a committee seems to be doing the writing. None of these ways is necessarily wrong.

For example, one approach might be: The original designer of the product writes the data sheet (inside his head) at the same time the product is designed. The concept here is, if one can't find the proper ingredients for a data sheet—good applications, convenient features for the user and nicely tested specifications as the part is being designed—then maybe it's not a very good product until all those ingredients are completed. Thus, the collection of raw materials for a good data sheet is an integral part of the design of a product. The actual assembly of these materials is an art which can take place later.

WHEN TO WRITE DATA SHEETS

A new product becomes available. The applications engineers start evaluating their application circuits and the test engineers examine their production test equipment.

But how can the users evaluate the new device? They have to have a data sheet—which is still in the process of being written. Every week, as the data sheet writer tries to polish and refine the incipient data sheet, other engineers are reporting, "These spec limits and conditions have to be revised," and, "Those application circuits don't work like we thought they would; we'll have one running in a couple of days." The marketing people insist that the data sheet must be finalized and frozen right away so that they can start printing copies to go out with evaluation samples.

These trying conditions may explain why data sheets always seem to have been thrown together under panic conditions and why they have so many rough spots. Users should be aware of the conflicting requirements: Getting a data sheet "as completely as possible" and "as accurately as possible" is compromised if one wants to get the data sheet "as quickly as possible."

The reader should always question the manufacturer. What are the alternatives? By not asking the right question, a misunderstanding could arise; getting angry with the manufacturer is not to anyone's advantage.

Robert Pease has been staff scientist at National Semiconductor Corp., Santa Clara, Calif., for eleven years. He has designed numerous op amps, data converters, voltage regulators and analog-circuit functions.

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Appendix G
Obsolete Product Replacement Guide

Some device types, individual temperature grades and package options have been discontinued. This guide is provided to help design engineers select and specify an appropriate alternative.

NSC Part Number	Replacement	Note
ADB1200	ADC3711	2
DAC1200/1201	DAC1265	2
LF352	LM3631	2
LF13300	ADC3711	2
LH0001	LM4250	2
LH0005/LH0005A	LH0003	2
LH0037	LH0036	3
LH0132	LH0032	2
LH2011	LM11	2
LH2201A	LM201A	2
LH2208	LM208	2
LH2208A	LM208A	2
LH24250	LM11	2
LM170/270/370	LM13600N	2
LM171/271/371	no replacement	
LM172/272/372	no replacement	
LM173/273/373	no replacement	
LM174/274/374	no replacement	
LM175/275/375	no replacement	
LM216/316	LM11	2
LM388N-2/N-3	LM388N-1	2
LM377N	LM2877P	3
LM378N	LM2878P	3
LM379	LM2879T	3
LM1014	no replacement	
LM1017	no replacement	
LM1019	no replacement	
LM1821S	LM1823	2

NSC Part Number	Replacement	Note
LM1822	LM1823	3
LM1828	no replacement	
LM1848	no replacement	
LM1877N-1/N-2/N-3	LM1877N-9	2
LM2003	no replacement	
LM2808	no replacement	
LM2831	LM1851	2
LM3011	no replacement	
LM3064	no replacement	
LM3075	no replacement	
TBA120V	no replacement	
TBA440C	LM1823	2
TBA510	no replacement	
TBA530	no replacement	
TBA540	no replacement	
TBA560C	no replacement	
TBA920	no replacement	
TBA950-2	no replacement	
TBA970	no replacement	
TBA990	no replacement	
TDA440	no replacement	
TDA2522/23	no replacement	
TDA2530	no replacement	
TDA2530/31	no replacement	
TDA2540/41	no replacement	
TDA2560	no replacement	
TDA2590	no replacement	
TDA3500	no replacement	

Note 1: IMPROVED REPLACEMENT: Pin for Pin replacement with superior electrical specifications.

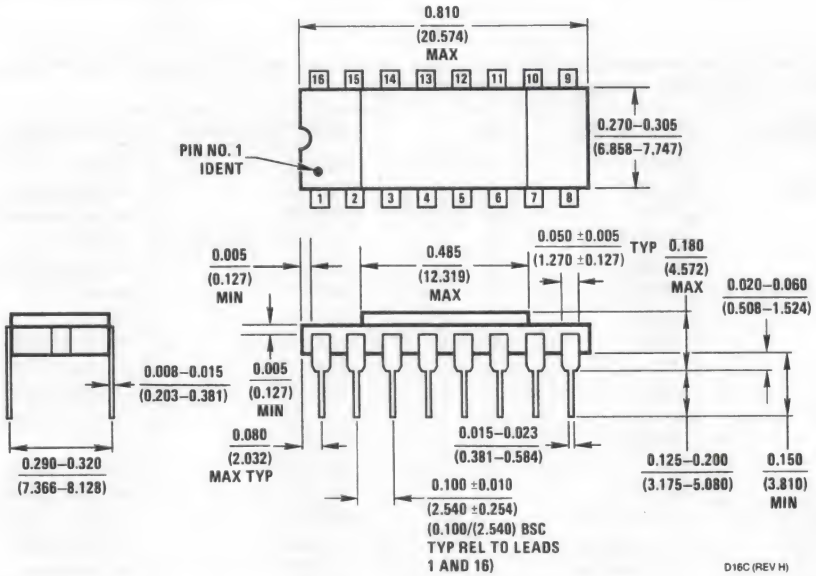
Note 2: FUNCTIONAL REPLACEMENT: Consult datasheet to determine suitability of the replacement for specific application.

Note 3: SIMILAR DEVICE with superior performance: Consult datasheet to determine suitability of the replacement for specific application.

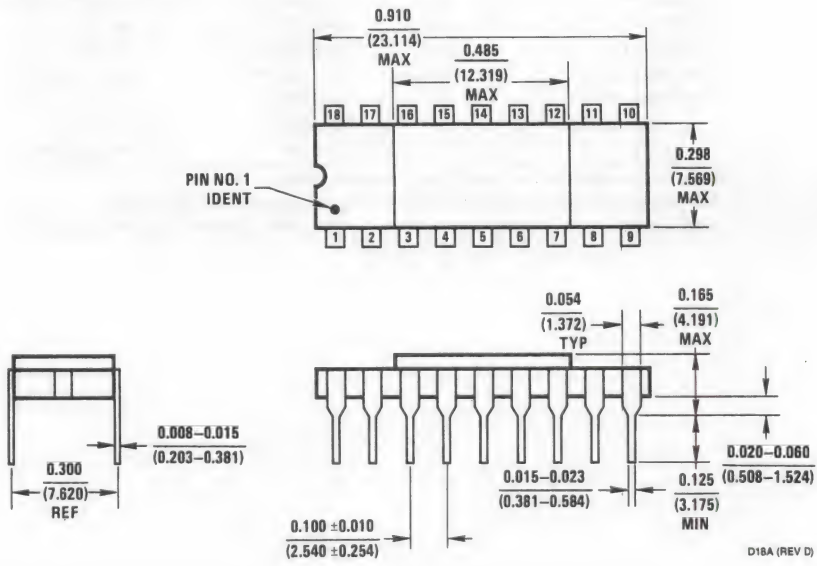
National Semiconductor

All dimensions are in inches (millimeters)

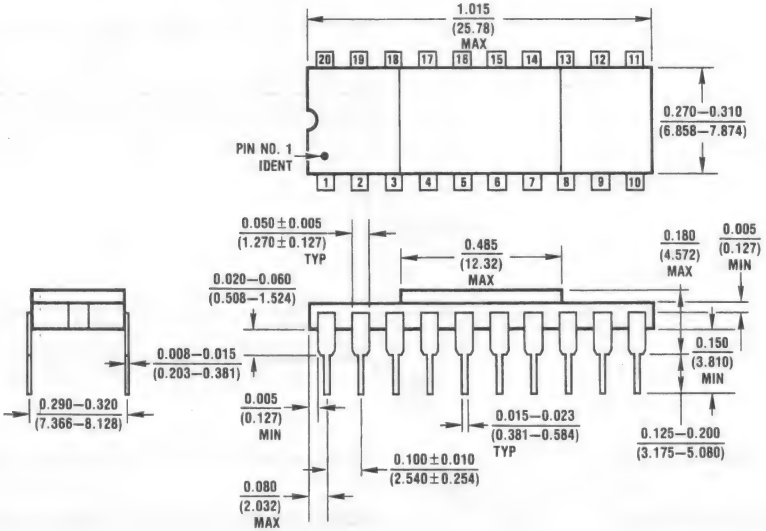
16 Lead Hermetic Dual-In-Line Package (D)
NS Package Number D16C



18 Lead Hermetic Dual-In-Line Package (D)
NS Package Number D18A

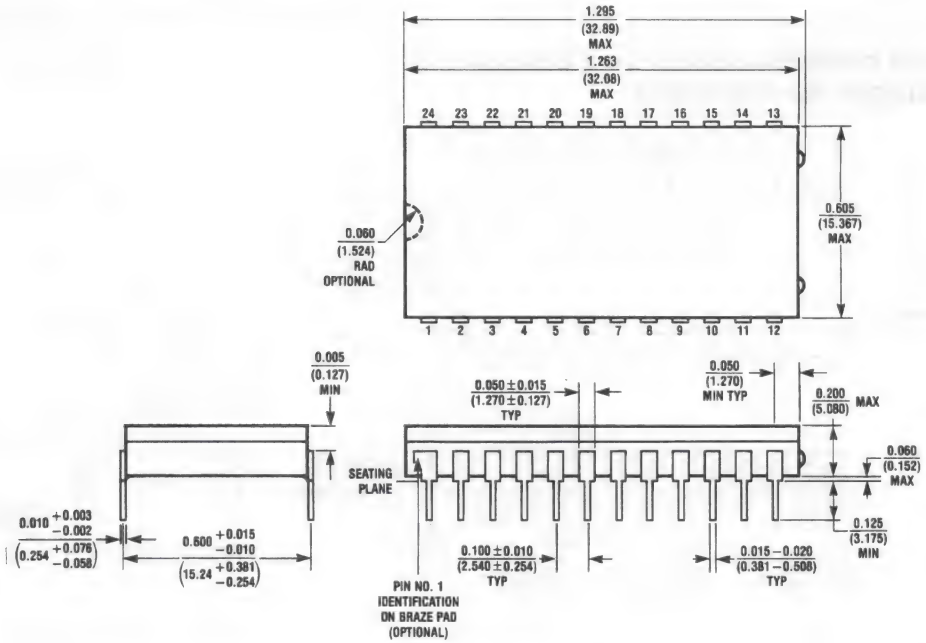


20 Lead Hermetic Dual-In-Line Package (D)
NS Package Number D20A



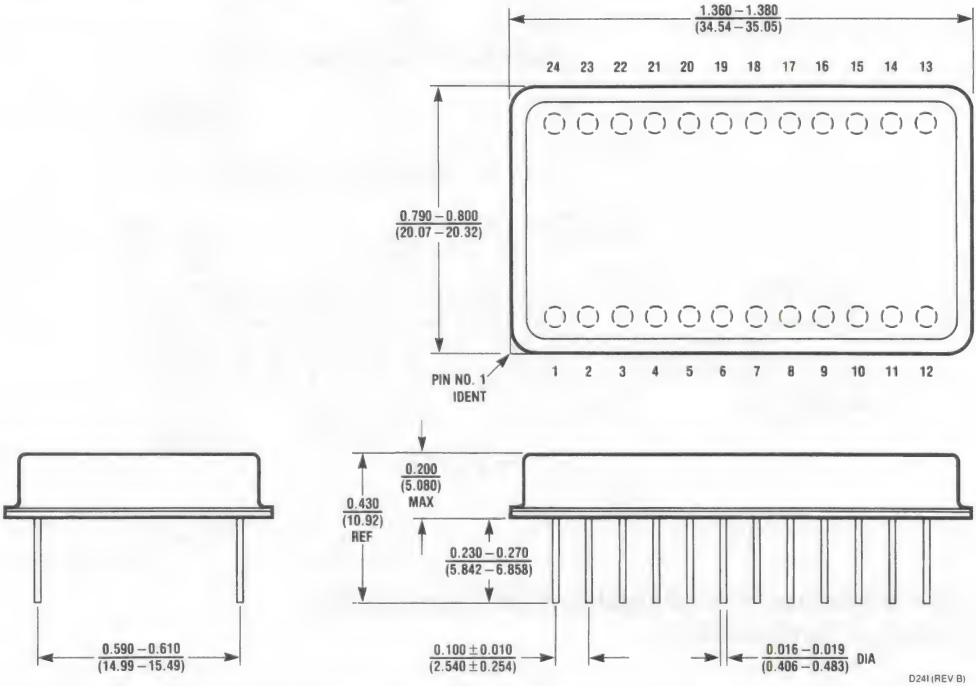
D20A (REV D)

24 Lead Sidebraze Hybrid Dual-In-Line Package (D)
NS Package Number D24D

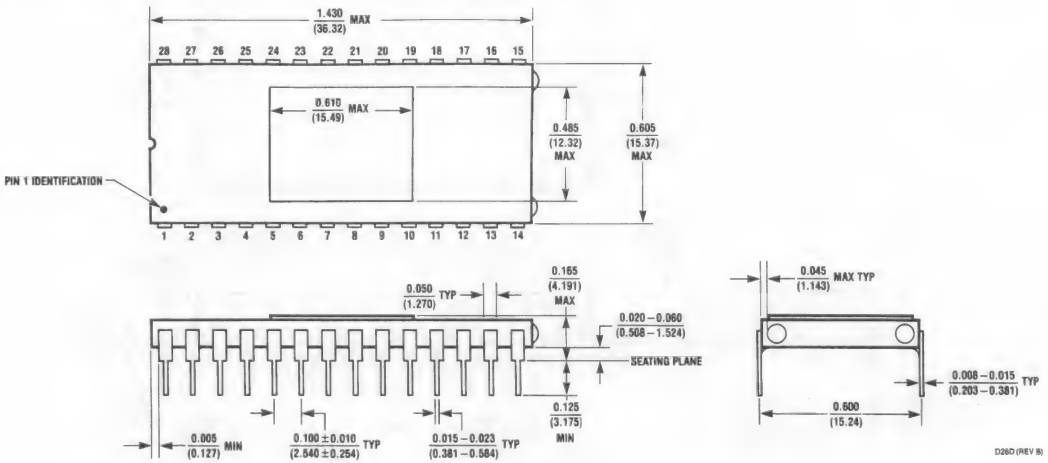


D24D (REV F)

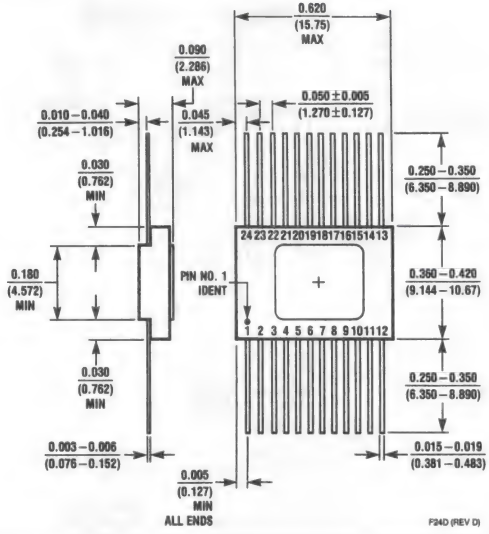
24 Lead (0.600" Wide) Dual-In-Line Metal Package (D)
NS Package Number D24I



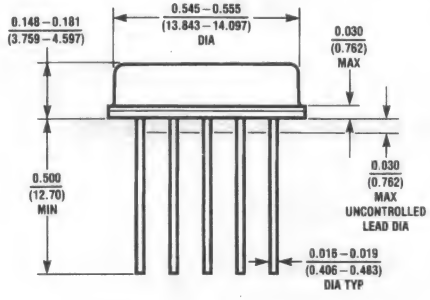
28 Lead Hermetic Dual-In-Line Package (D)
NS Package Number D28D

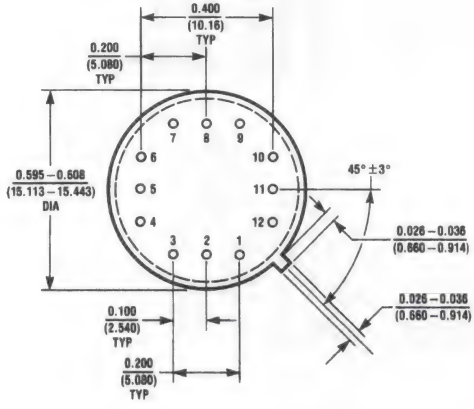


24 Lead Ceramic Flatpack (F)
NS Package Number F24D

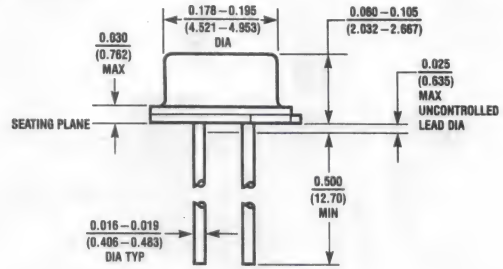


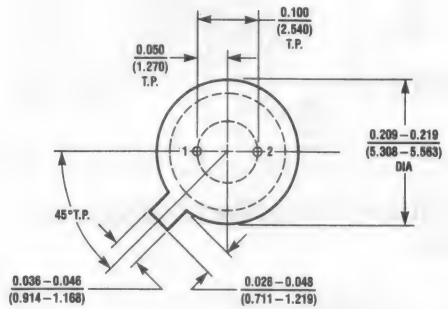
12 Lead (0.400″ Square Pattern) Metal Can Package (G)
NS Package Number G12B





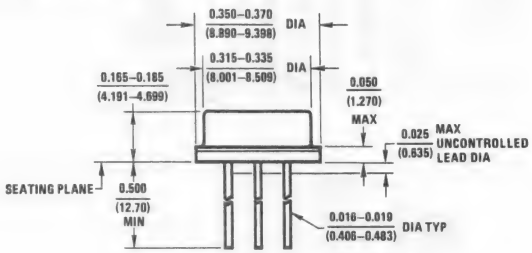
2 Lead (0.100″ Diameter P.C.) TO-46 Metal Can Package (H)
NS Package Number H02A

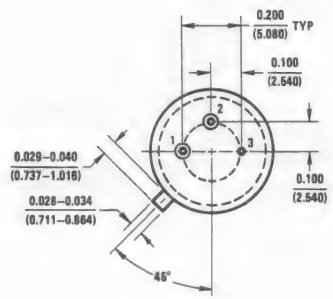




H02A (REV C)

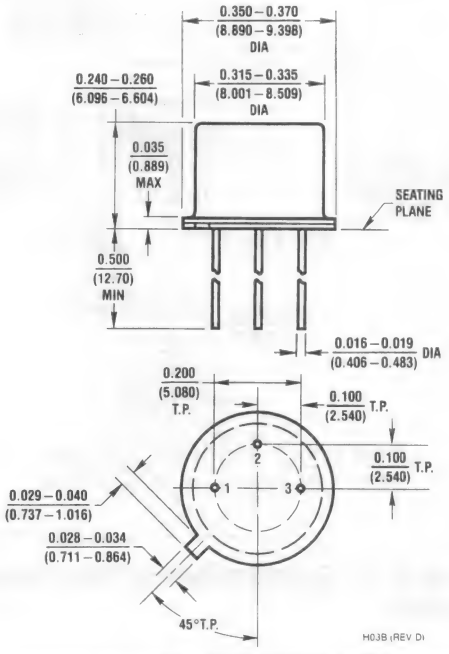
3 Lead (0.200″ Diameter P.C.) Metal Can Package (H)
NS Package Number H03A



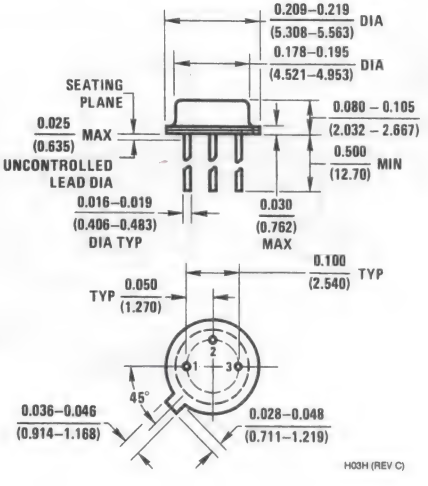


H03A (REV B)

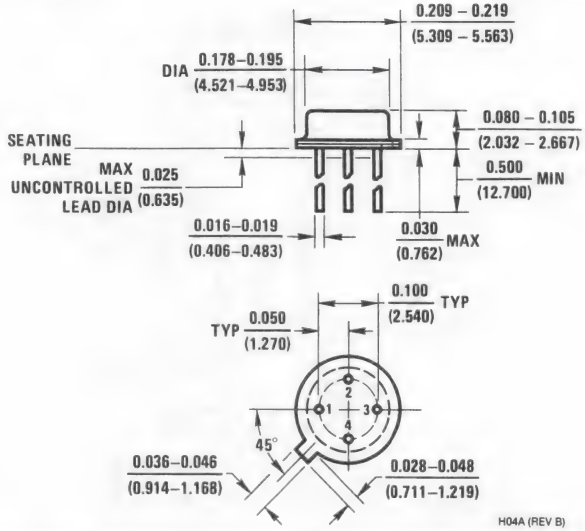
3 Lead (0.200″ Diameter P.C.) TO-39 Metal Can Package (H)
NS Package Number H03B



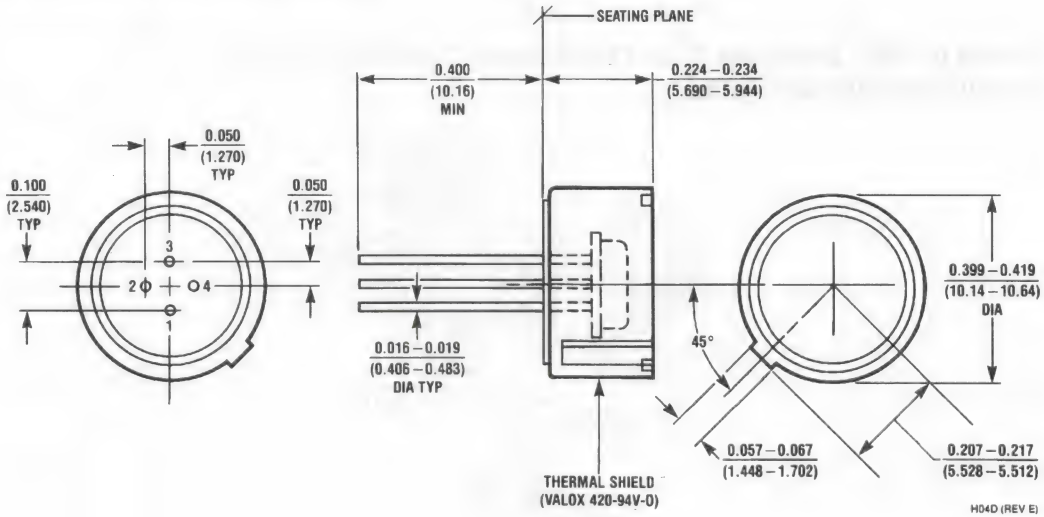
3 Lead (0.100″ Diameter P.C.) TO-46 Metal Can Package (H)
NS Package Number H03H



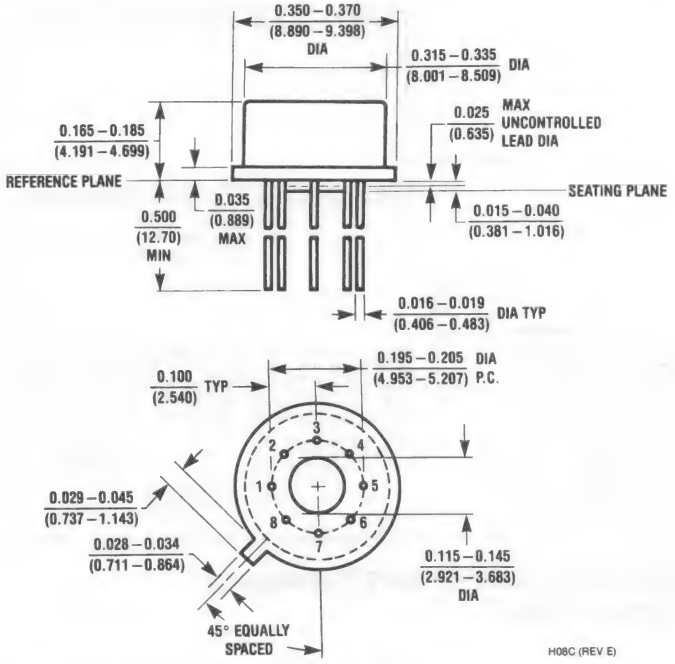
4 Lead (0.100″ Diameter P.C.) TO-46 Metal Can Package (H)
NS Package Number H04A



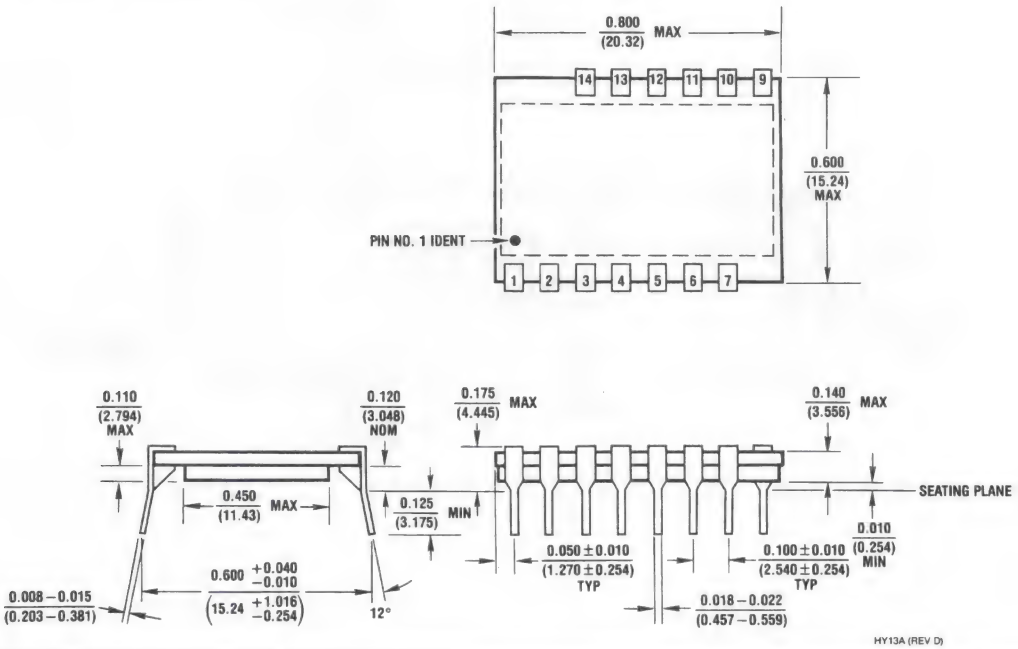
4 Lead (0.100″ Diameter P.C.) Shielded Metal Can Package (H)
NS Package Number H04D



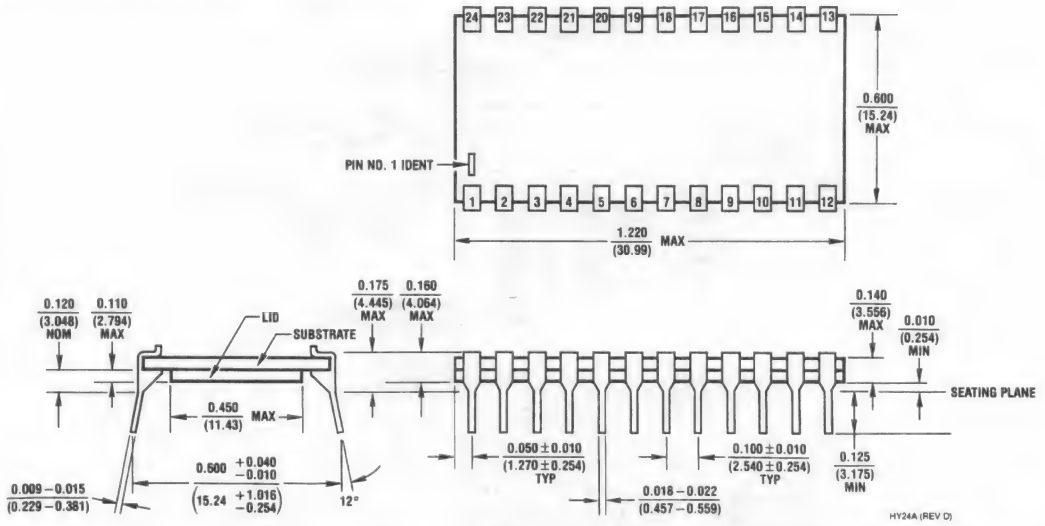
8 Lead (0.200″ Diameter P.C.) TO-99 Metal Can Package (H)
NS Package Number H08C



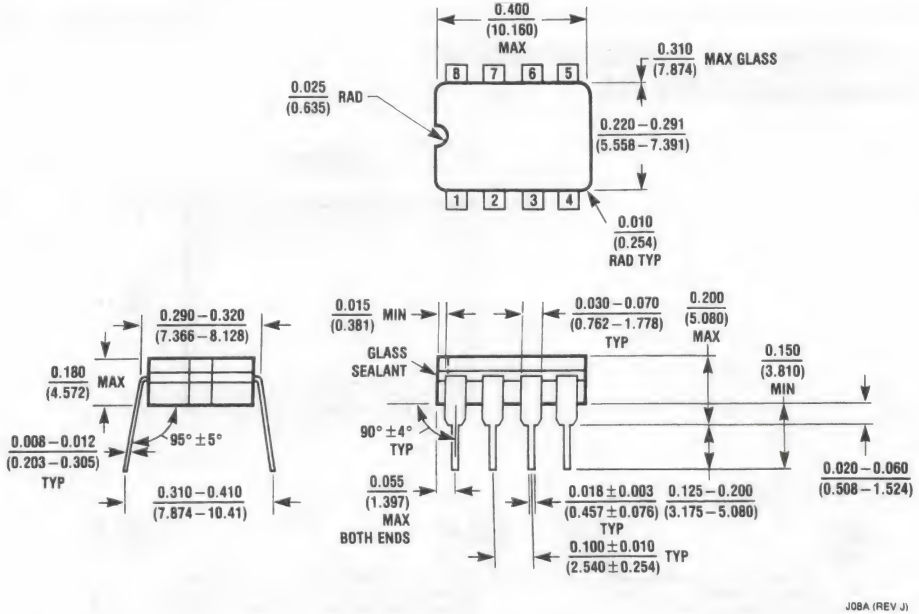
13 Lead Dual-In-Line Hybrid Package (J)
NS Package Number HY13A



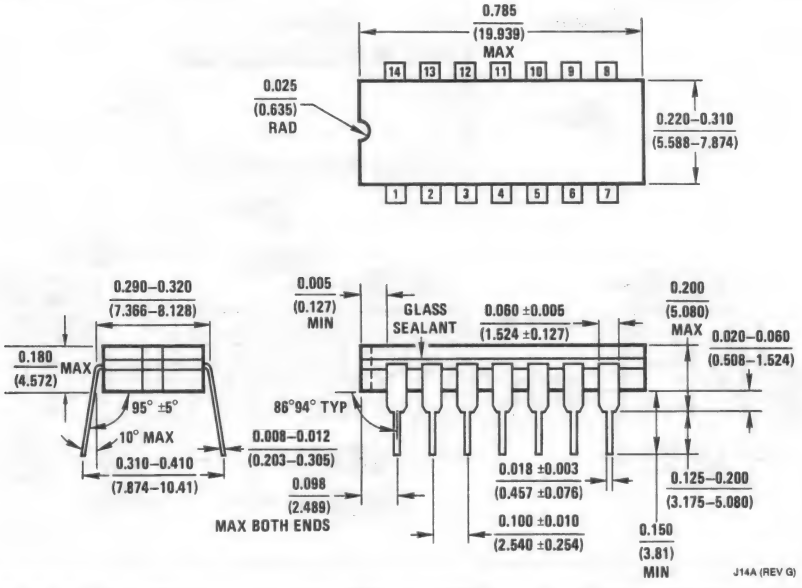
24 Lead Dual-In-Line Hybrid Package (J)
NS Package Number HY24A



8 Narrow Lead Ceramic Dual-In-Line Package (J)
NS Package Number J08A



14 Lead Ceramic Dual-In-Line Package (J)
NS Package Number J14A



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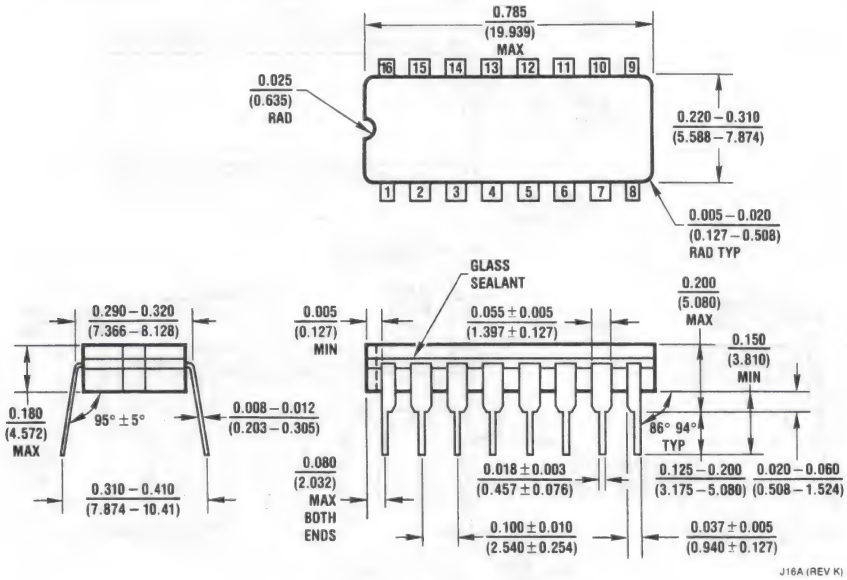

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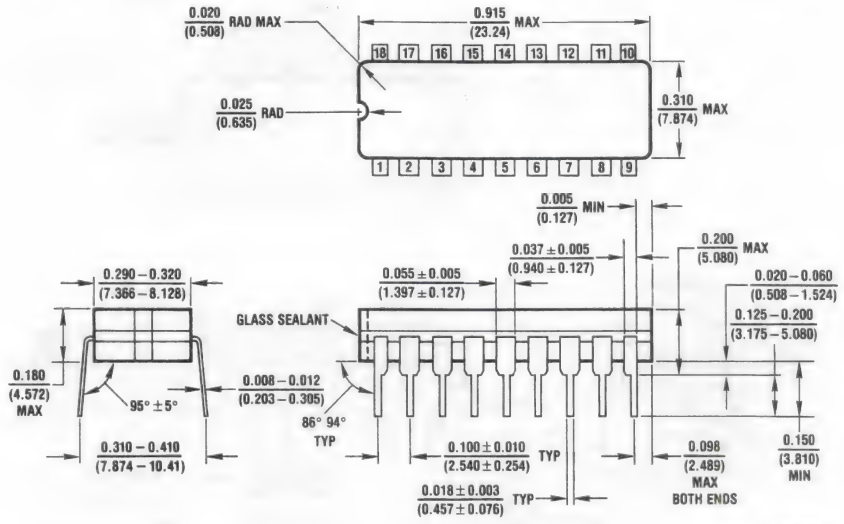

J18A (REV L)

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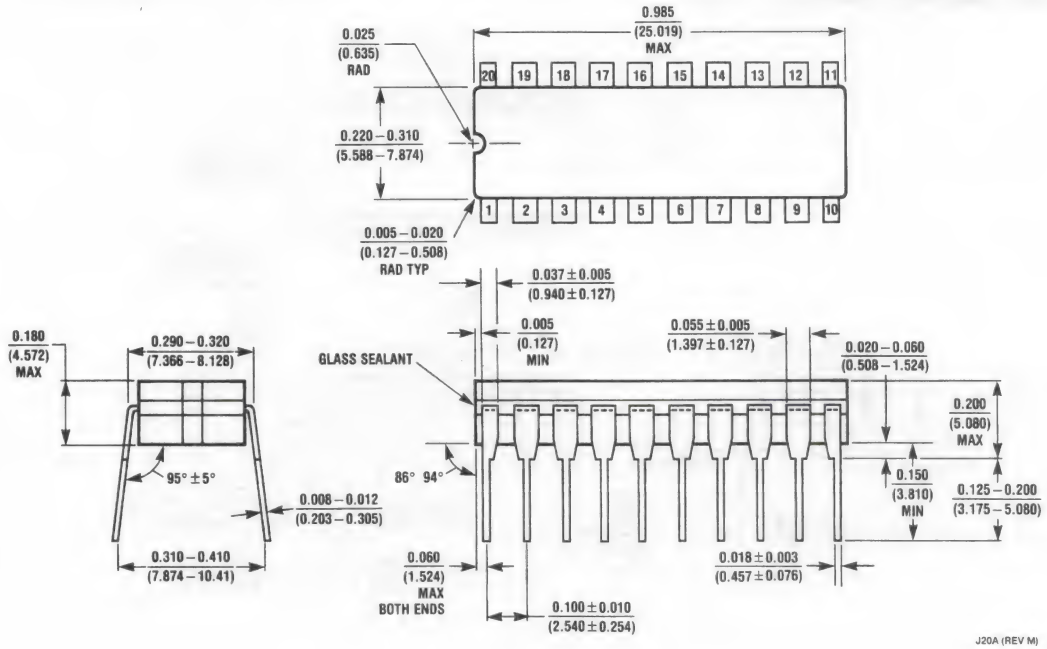

J20A (REV M)

24 Lead Ceramic Dual-In-Line Package (J)
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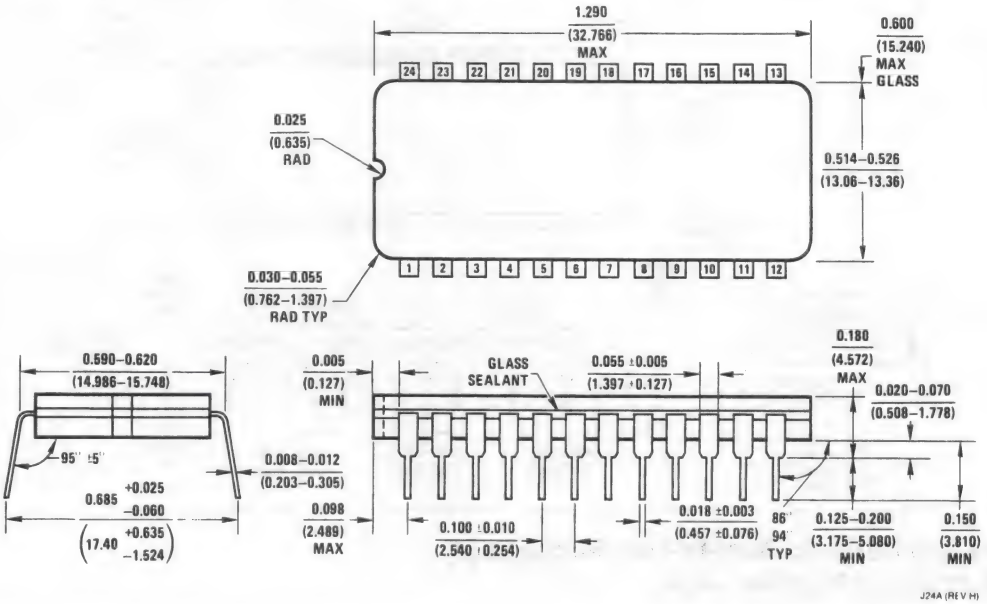

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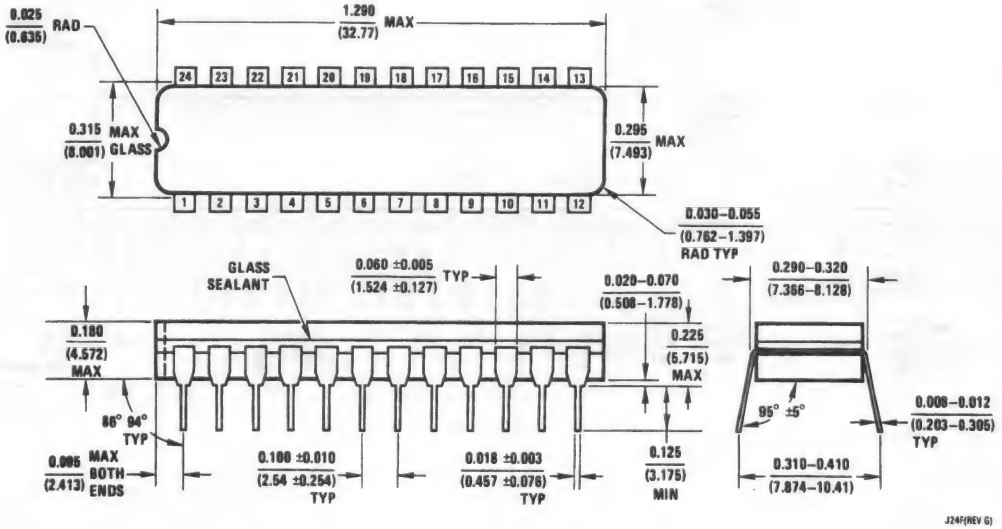
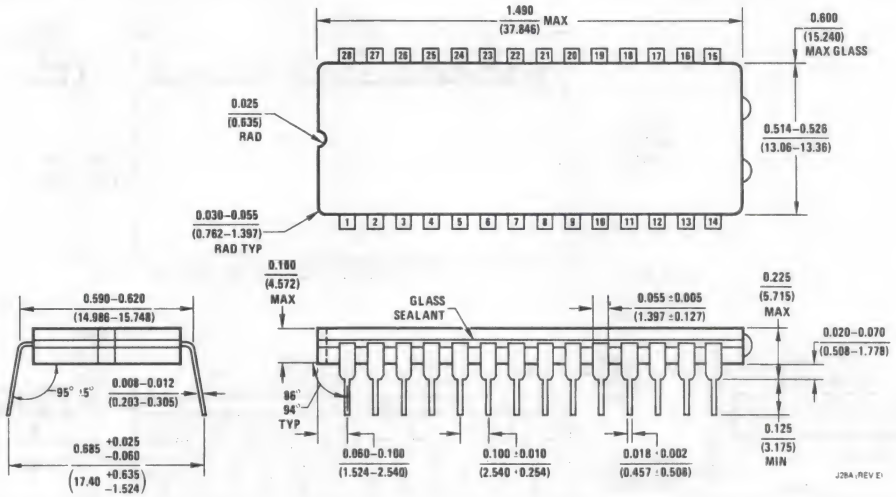


J24F(REV G)

28 Lead Ceramic Dual-In-Line Package (J)
NS Package Number J28A



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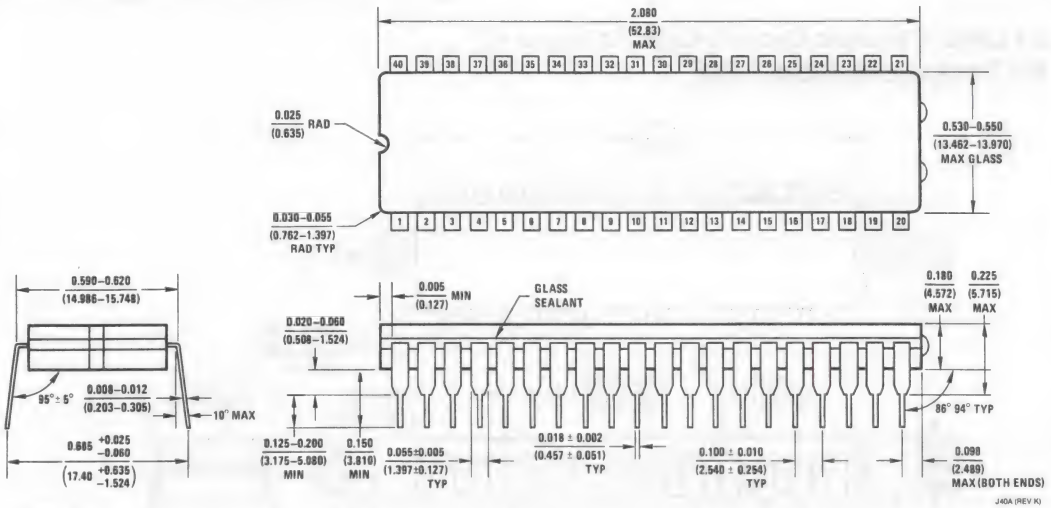
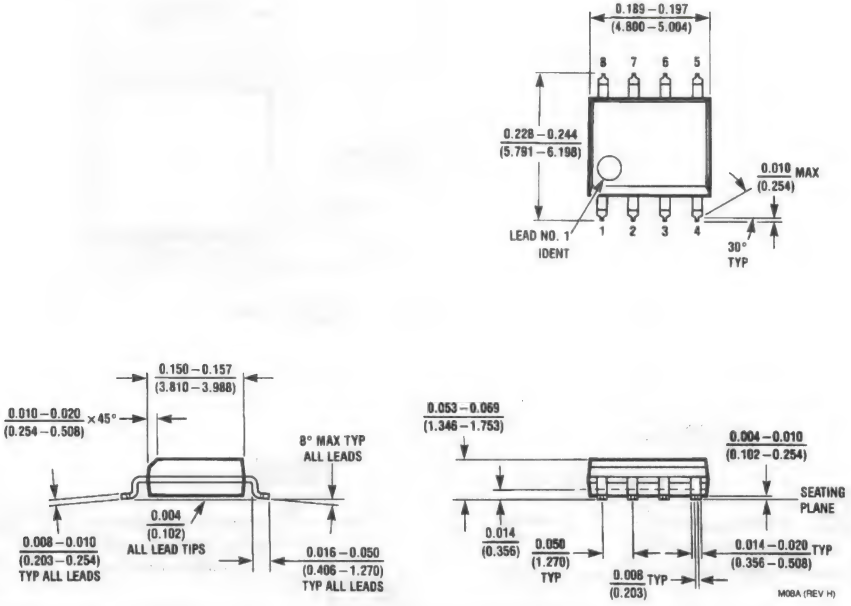
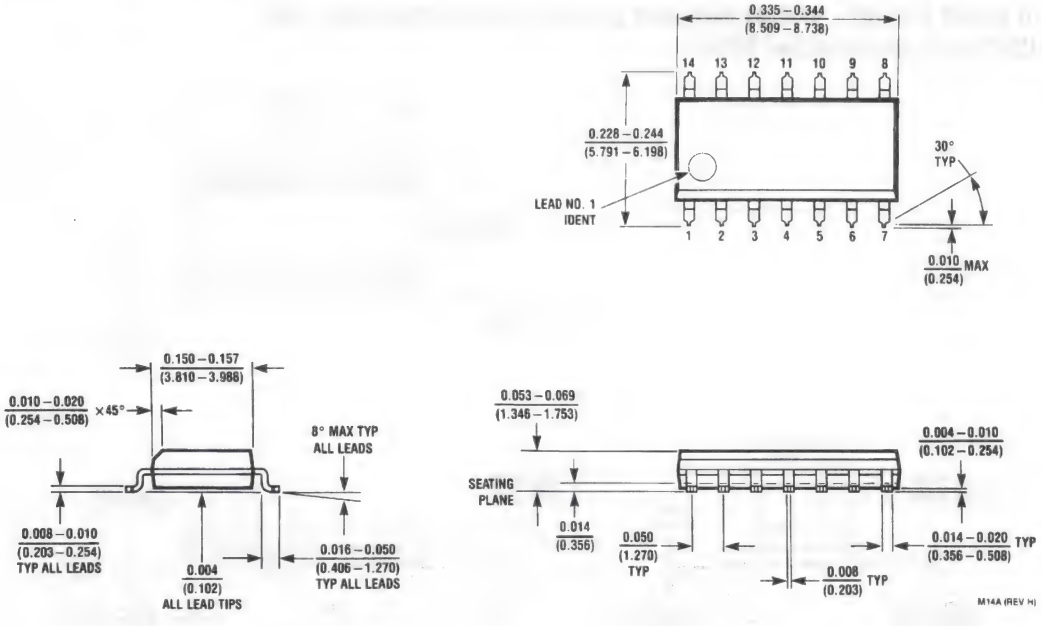


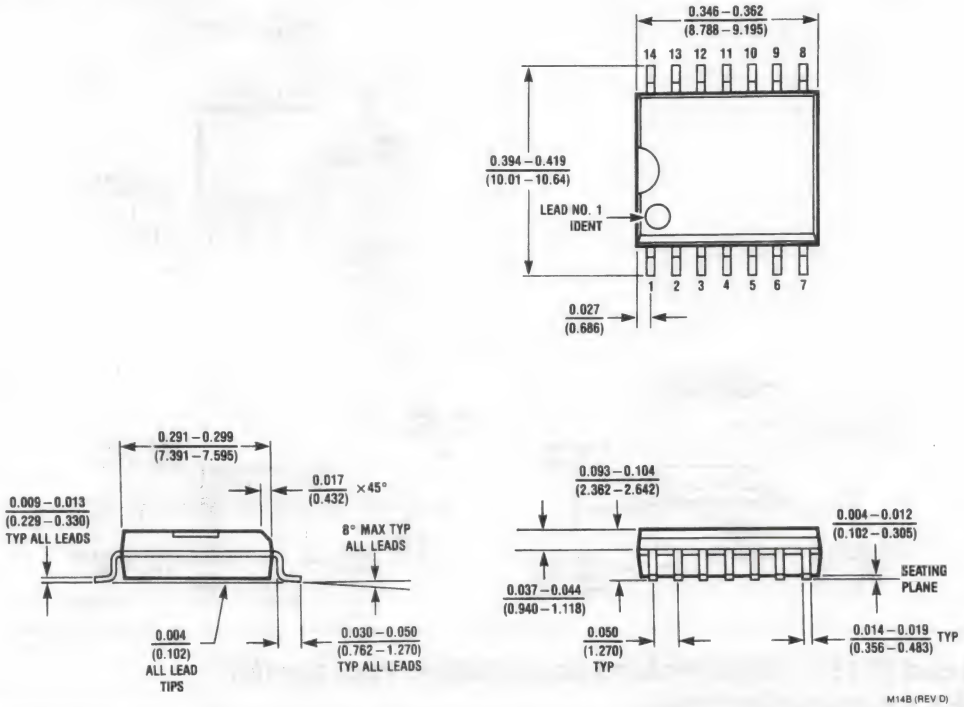
8 Lead (0.150″ Wide) Molded Small Outline Package (M)
NS Package Number M08A



14 Lead (0.150″ Wide) Molded Small Outline Package (M)
NS Package Number M14A



14 Lead (0.300" Wide) Molded Small Outline Package (M)
NS Package Number M14B



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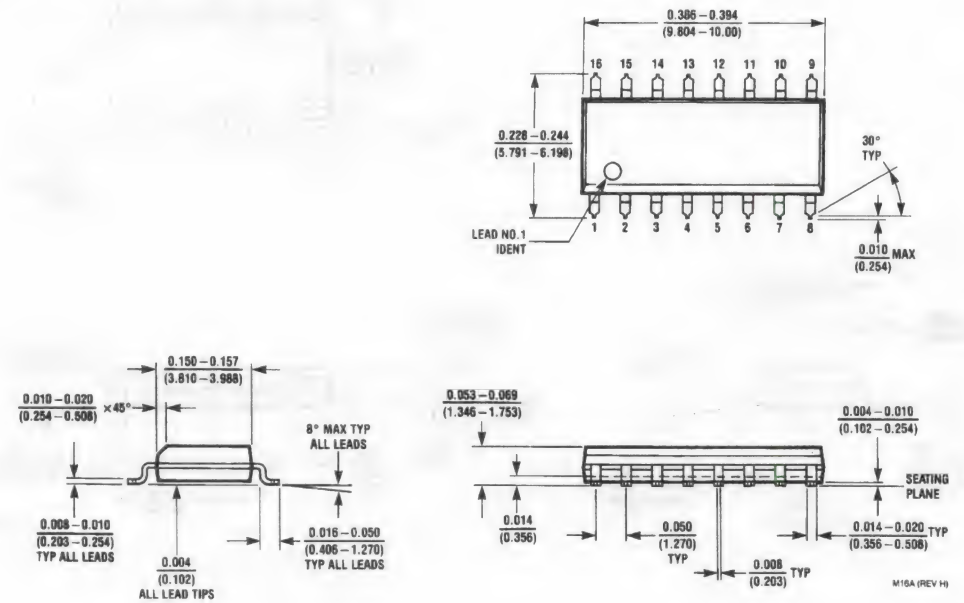
NS Package Number M16A



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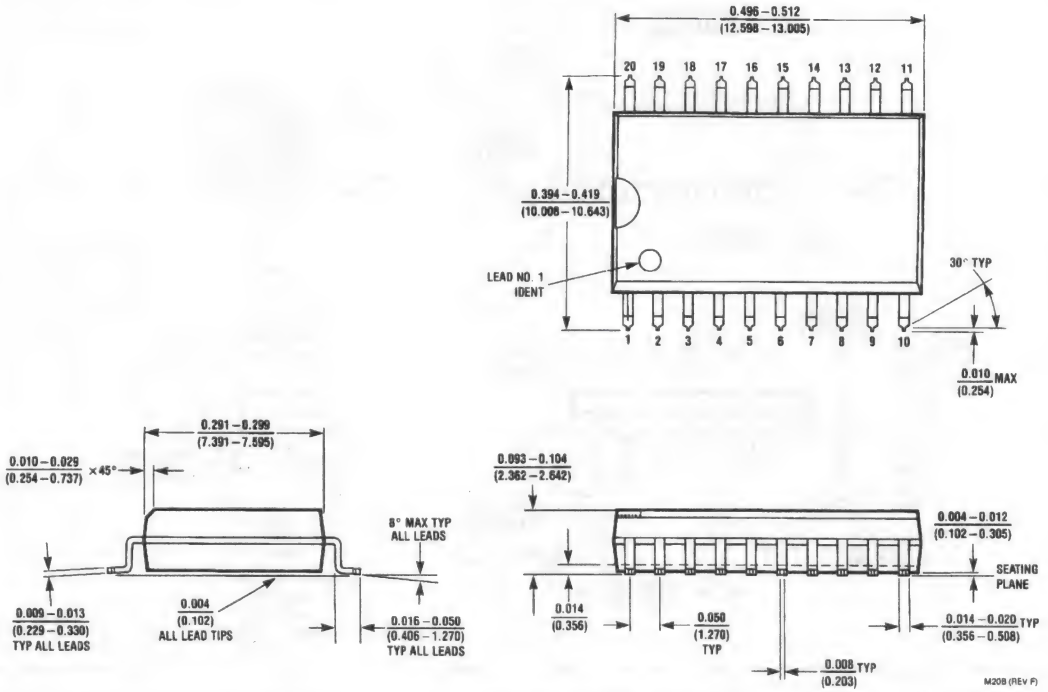

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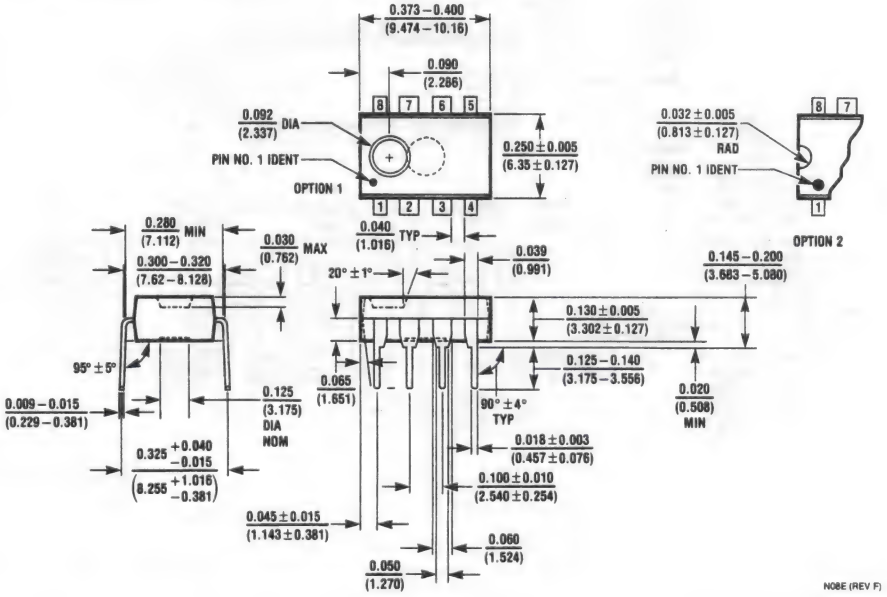
NS Package Number N08E



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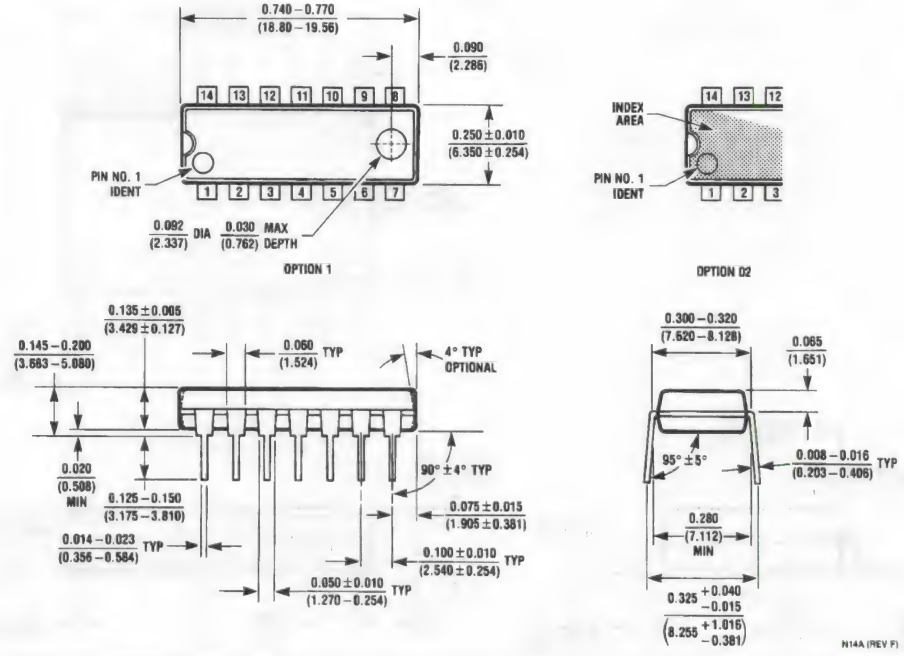
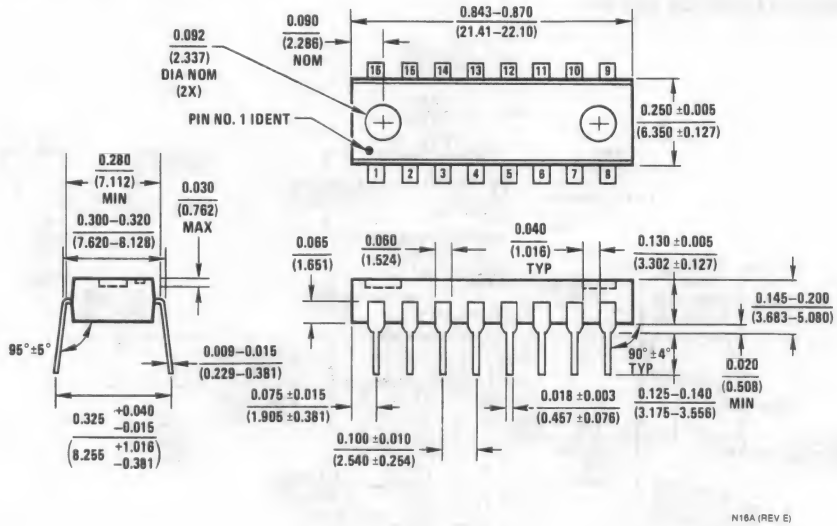
NS Package Number N14A



16 Lead Molded Dual-In-Line Package (N)
NS Package Number N16A



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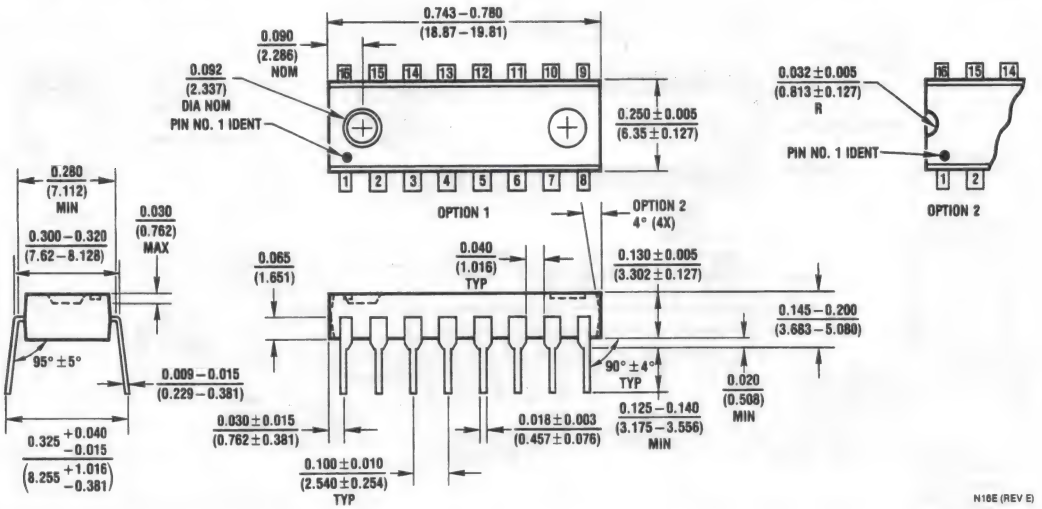

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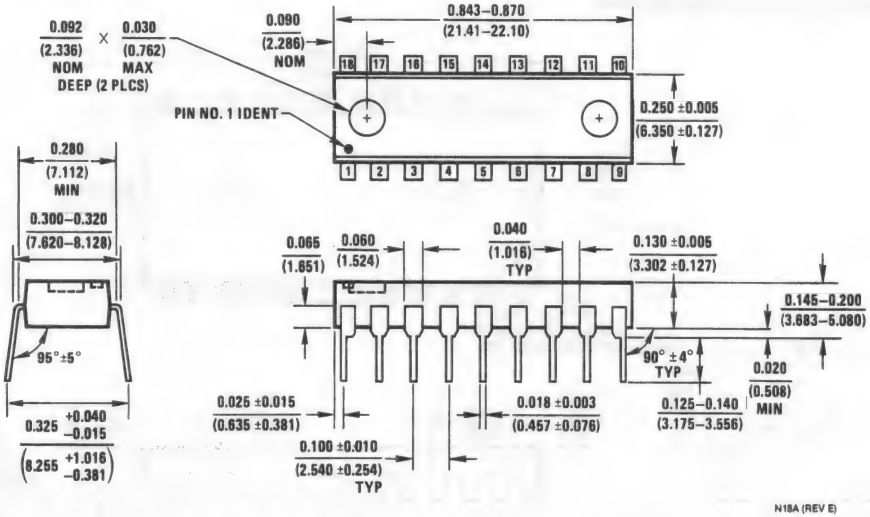

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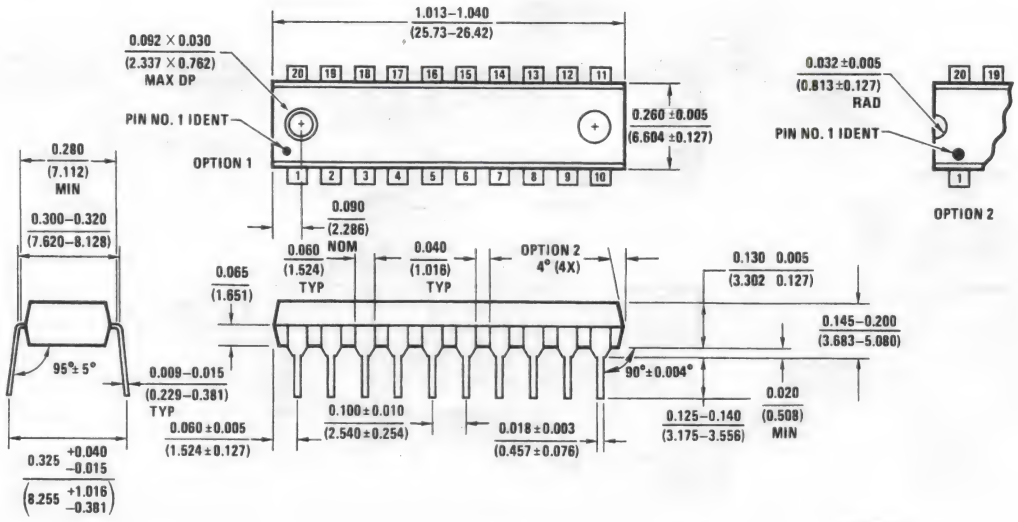

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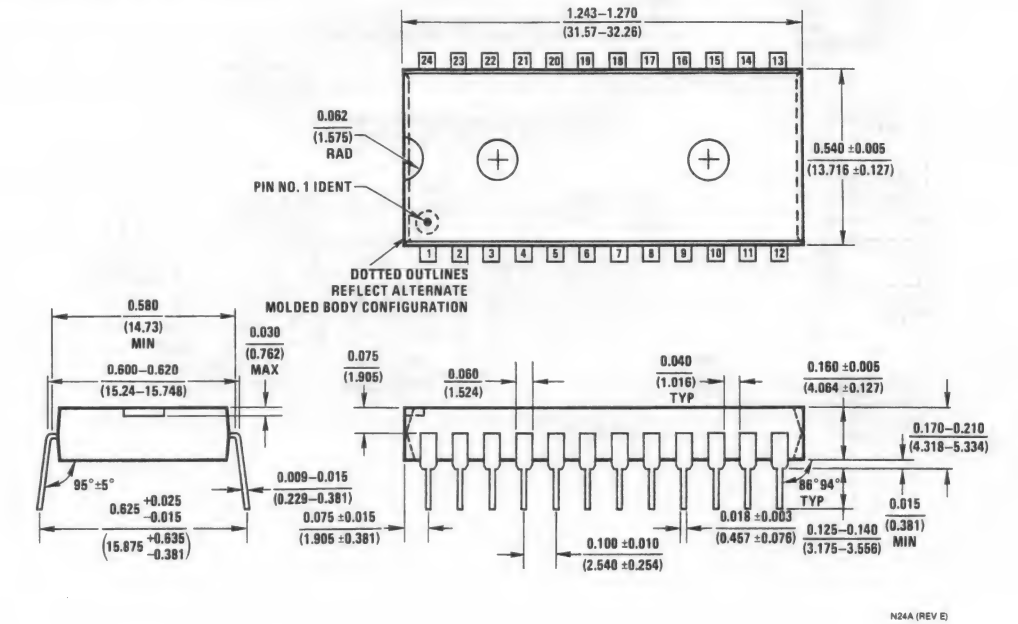
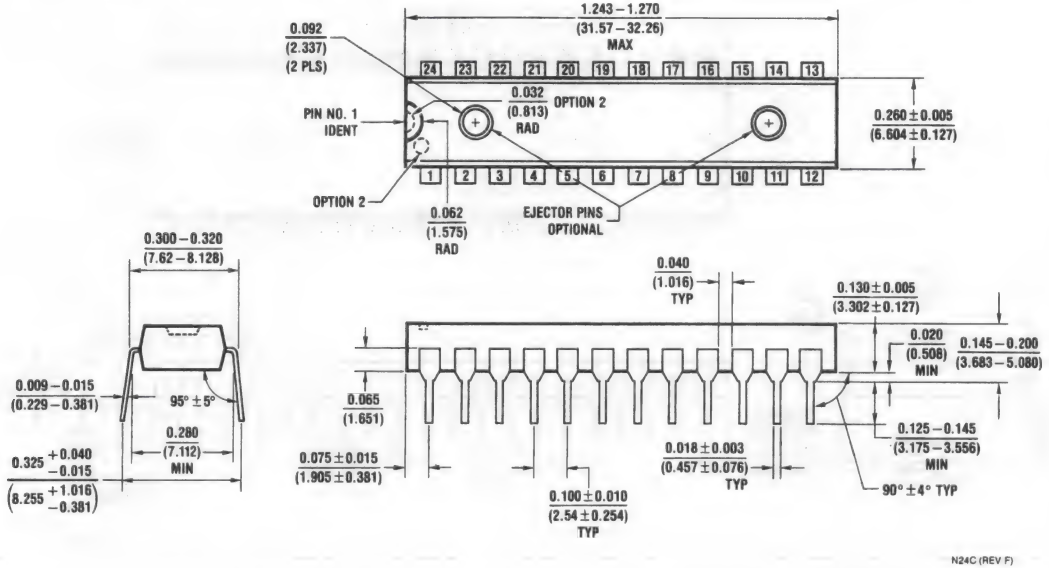
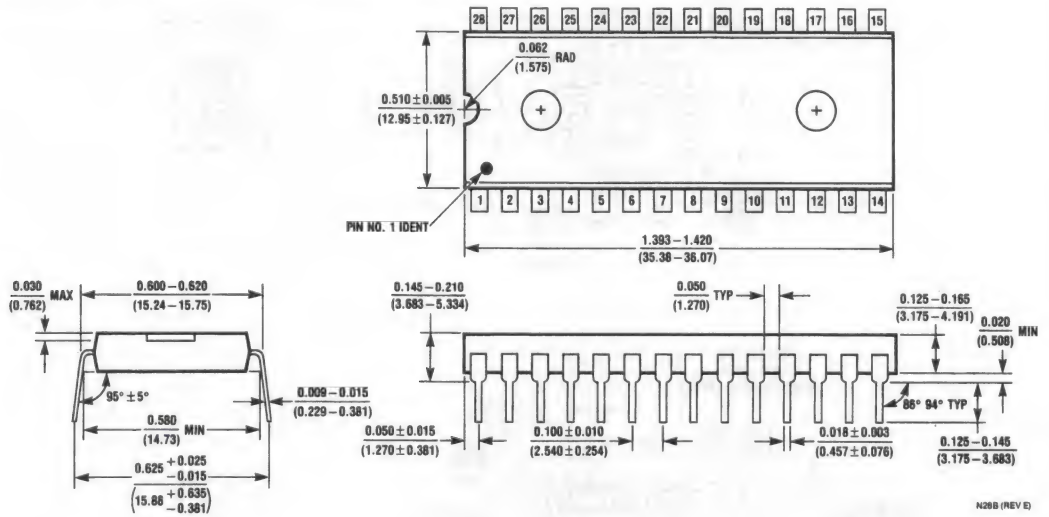


24 Lead Skinny Dual-In-Line Package (0.300″ Centers Molded) (N)
NS Package Number N24C



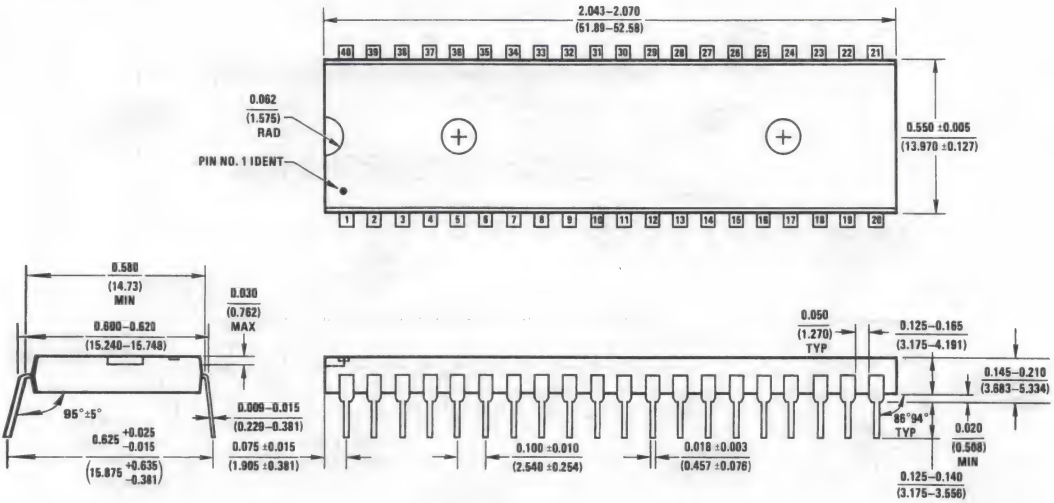
N24C (REV F)

28 Lead Molded Dual-In-Line Package (N)
NS Package Number N28B



N28B (REV E)

40 Lead Molded Dual-In-Line Package (N)
NS Package Number N40A



N40A (REV E)

20 Lead Plastic Chip Carrier (V)
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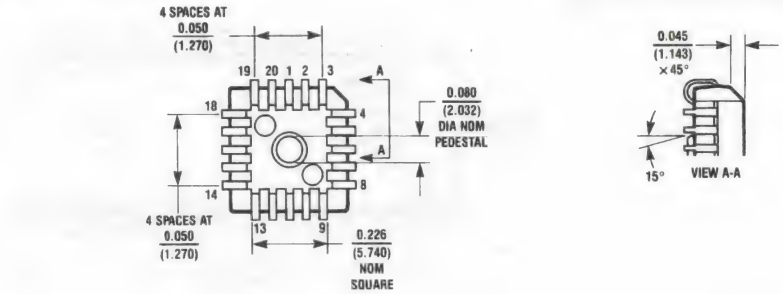

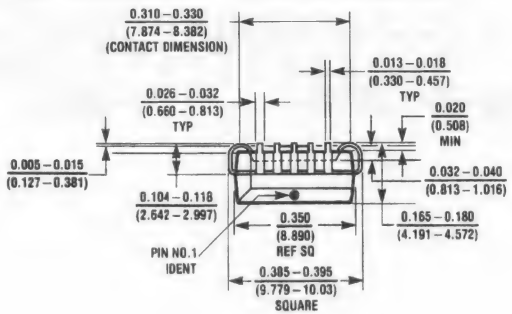

V20A (REV J)

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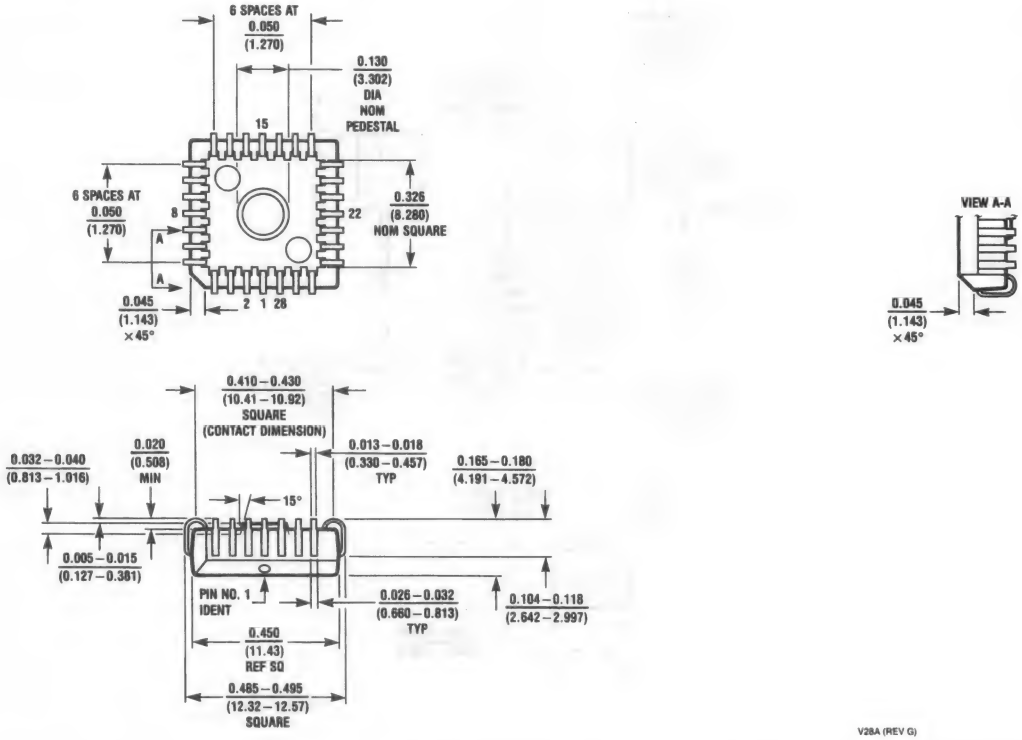
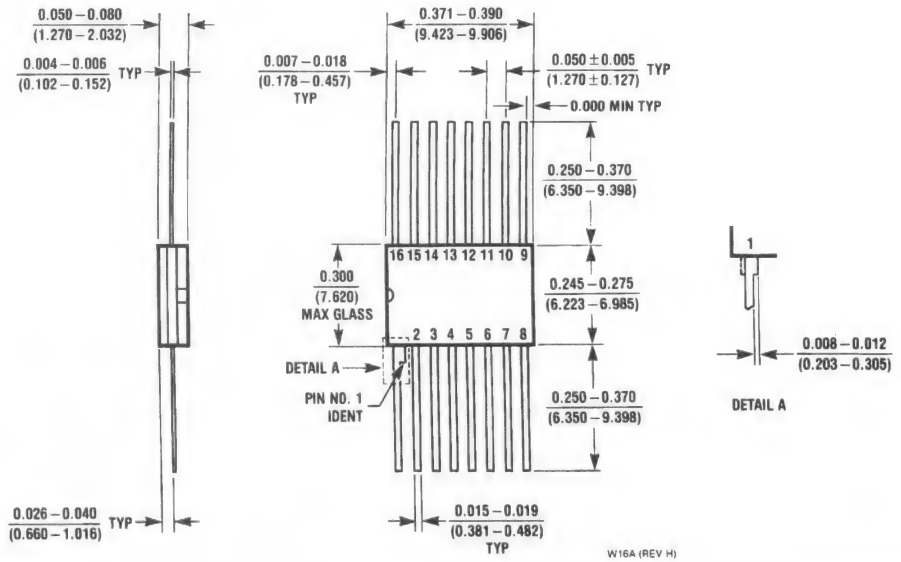
NS Package Number V28A



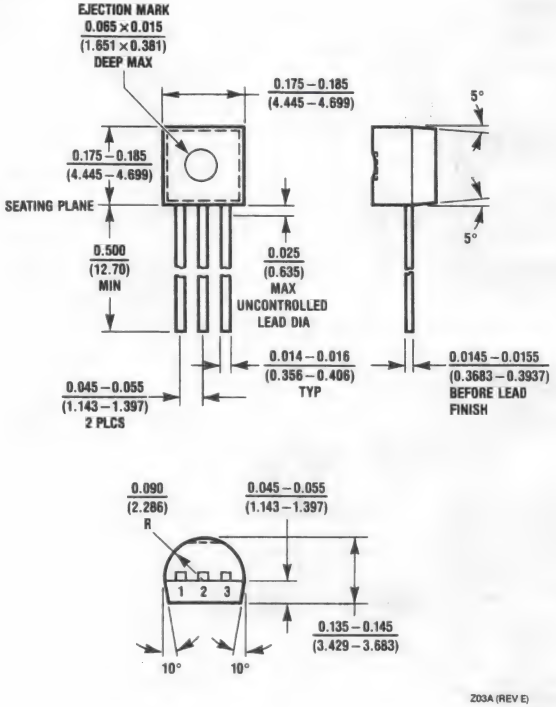
V28A (REV G)

16 Lead Ceramic Package (W)
NS Package Number W16A



W16A (REV H)

3 Lead TO-92 Molded Package (Z)
NS Package Number Z03A



Z03A (REV E)

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